SUBCOURSE MM0323 | EDITION 6

FM RADIO TRANSMITTERS

AIPD

READINESS/ PROFESSIONALISM THRU GROWTH

THE ARMY INSTITUTE FOR PROFESSIONAL DEVELOPMENT

ARMY CORRESPONDENCE COURSE PROGRAM

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*May 1971

CORRECTIONS TO TM 11-668

Page 9, para 4d(4), lines 30 and 31. Change the formula to:

$$\underline{\Delta F} = \frac{\pi}{6} \times 1{,}000 \times (+1)$$

$$\underline{\Delta F} = +523 \text{ cps (approximately).}$$

Page 20, para 11g(2), line 5. Change "25" to: 15.

Page 29, para 13d, make the following changes:

Line 16. Change "750,000 to: 75,000.

Line 18. Change "750,000" and ".001" to: 75,000 and $.001 \times 10^{-6}$, respectively.

Page 36, para 18c, line 1. Change "shunt-fed" to: series-fed.

Figure 36, caption. Change "Shunt fed" to: Series-fed.

Page 37, para 20, first formula. Change to:

$$X_C = \frac{1}{2\pi f\underline{C}} \text{ ohms.}$$

Page 40, para 22b(3), formula at top of page. Change to:

$$Z_{ab} = \frac{1}{g_m} \times \left(\frac{Z_a + Z_b}{Z_b}\right)$$

$$Z_{ab} = \frac{1}{g_m} \times \left(1 + \frac{Z_a}{Z_b}\right) \qquad g_m = \text{mhos}$$

$$Z_{ab} = \frac{1}{g_m} + \frac{Z_a}{g_m Z_b}$$

Page 94, para 43d(3). Delete line 7, and substitute: less; therefore, a positive voltage.

Page 118, para 59a, lines 17-25. Change to: The over-all output is the quadrature sum of the signal and the noise voltages, multiplied by the stage amplification, or

$$\sqrt{10^2 + 4.4^2} \times 10 = \underline{10.9} \times 10 = 109 \text{ microvolts.}$$

*This edition replaces correction sheet dated April 1969.

Since the second stage of amplification was assumed to be identical with the first, it add 3.2 microvolts of noise to the applied signal of 109 microvolts. The output of the second stage is

$$\sqrt{109^2 + 3.2^2} \times 10 = \underline{109} \times 10 = \underline{1{,}090 \text{ microvolts}}.$$

Page 151, figure 134 A and B. In the label on the vertical side of each graph, change "RESPONSE" to: AMPLITUDE.

PLEASE NOTE

Proponency for this subcourse has changed from Signal (SS) to Missile & Munitions'(MM).

SIGNAL SUBCOURSE 323, FM RADIO TRANSMITTERS

INTRODUCTION

In a field army, a sufficient number of radio sets are available to provide radio communications for all commanders. Most of these radio sets use a type of modulation that is called frequency modulation (FM). These FM radio sets are used to pass various types of communication traffic and also appear in various sizes and shapes. For example: these sets vary from the relatively simple squad type to the highly sophisticated type used in satellite communication terminals. Even though these sets transmit FM signals, there are different methods used to develop these modulated signals.

This subcourse is written to make you aware of different frequency-modulating techniques and to acquaint you with various types of circuits used in FM transmitters.

This subcourse consists of four lessons as follows:

Lesson 1. Principles of FM

Lesson 2. Methods of Producing FM

Lesson 3. FM Transmitter Circuits

Lesson 4. FM Transmitters

Credit Hours: 10

You are urged to finish this subcourse without delay; however, there is no specific limitation on the time you may spend on any lesson or on the examination.

Texts and materials furnished:

Subcourse Booklet

TM 11-668, F-M Transmitters and Receivers, September 1952

REVIEWED AND REPRINTED WITH MINOR REVISIONS AUGUST 1976

Note to student:

Since TM 11-668 was published, the unit of frequency has been changed from cycles per second to hertz. The following equivalents will help you relate the two units.

1 cycle per second (cps) = 1 hertz (Hz)

1 kilocycle per second (kcs) = 1 kilohertz (kHz)
(sometimes kc)

1 megacycle per second (mcs) = 1 megahertz (MHz)
(sometimes mc)

LESSON 1

PRINCIPLES OF FM

SCOPE ....Principles of amplitude modulation (AM), phase modulation (PM), and frequency modulation (FM); effects of the modulating signal's amplitude and frequency on the carrier wave; bandwidth occupied by an FM signal.

CREDIT HOURS 2

TEXT ASSIGNMENT ....TM 11-668, para 1-15; Attached Memorandum, para 1-1

MATERIALS REQUIRED ....None

SUGGESTIONS ....None

LESSON OBJECTIVES

When you have completed this lesson, you will be able to:

1. Compute the needed carrier amplitude and frequency for an FM transmitter when given the limiting conditions.

2. Determine the modulation index and the carrier deviation for an FM transmitter.

3. Determine the changes in signal bandwidth when the modulation index is changed.

4. Use the time-constant formulas to determine circuit response and the component values that are needed for a particular response.

ATTACHED MEMORANDUM

1-1. DEFINITIONS

The following definitions will be used in this lesson.

a. Frequency Deviation. The amount by which a frequency-modulated wave either increases or decreases from the center frequency.

b. Frequency Swing. The peak difference between the maximum and minimum values of instantaneous frequency. Normally considered as twice the frequency deviation.

LESSON EXERCISES

In each of the following exercises, select the ONE answer that BEST completes the statement or answers the question. Indicate your solution by circling the letter opposite the correct answers in the subcourse booklet.

1. The waveform shown in figure 1-1 represents a portion of a carrier wave being radiated from a radio transmitter. The frequency and peak voltage outputs of this transmitter, respectively, are

a. 80 kHz and 5 volts.

b. 80 kHz and 10 volts.

c. 160 kHz and 5 volts.

d. 160 kHz and 10 volts.

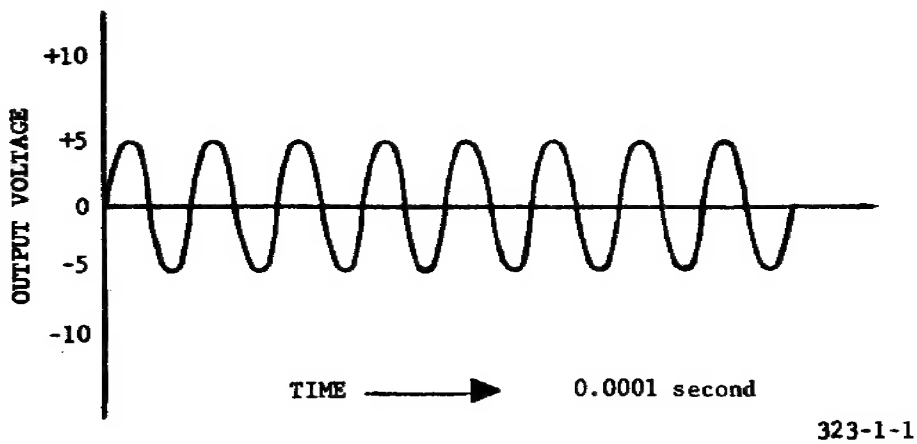


Figure 1-1. Unmodulated carrier wave.

2. Energy can be transmitted over great distances via a radio-frequency carrier wave. Intelligence can be "carried" by this wave if any one of its three characteristics is varied. These characteristics are

a. frequency, phase, and amplitude.

b. starting point, phase, and frequency.

c. amplitude, frequency, and starting point.

d. phase, amplitude, and signal-to-noise ratio.

3. To prevent distortion of the output signal, the operator of an AM radio must keep modulation from going over 100 percent. At the same time, modulation must be as near 100 percent as possible to obtain the maximum power output. Since the modulating signal may reach a peak amplitude of 30 volts, the carrier must have a peak amplitude of at least

a. 15 volts.

b. 30 volts.

c. 45 volts.

d. 60 volts.

4. A commercial AM radio station is transmitting a carrier frequency of 1,120 kHz. The frequency of the modulating signals ranges from 500 Hz to 5 kHz. The radio-frequency (RF) stages of the receiver being used to receive these signals must have a minimum bandwidth of approximately

a. 5 kHz.

b. 10 kHz.

c. 560 kHz.

d. 1,125 kHz.

5. Assume that a phase modulator is producing a 100-kHz carrier. The phase deviation limit is 30° when the carrier is modulated by a 1.2-kHz signal. The minimum and maximum frequency values of the modulated carrier are approximately

a. zero and 628 Hz.

b. 99,372 Hz and 100,628 Hz.

c. 99,686 Hz and 100,314 Hz.

d. 100,000 Hz and 100,628 Hz.

6. Assume that two identical sine-wave sources are used to modulate two transmitters--one is FM and the other is phase modulated (PM). The relationship between the frequency shifts in the two waves is such that the FM wave is at

a. maximum frequency when the PM wave is at its carrier frequency.

b. minimum frequency when the PM wave is at its maximum frequency.

c. minimum frequency when the PM wave is at its minimum frequency.

d. carrier frequency when the PM wave is at its carrier frequency.

7. Sketch C in figure 17 of TM 11-668 is the block diagram of a PM transmitter. The section that is relied upon to produce the maximum amount of frequency deviation is the

a. power amplifier.

b. crystal oscillator.

c. frequency multipliers.

d. audio-correction network.

8. Most military FM transmitters are restricted to maximum frequency deviation of 40 kHz.. If the maximum audio signal is 5 kHz, the modulation index is

a. 1/8.
b. 1/4.
c. 2.
d. 8.

9. Assume an FM station with a maximum carrier deviation of 40 kHz. If the modulation index (MI) is to be 5, the maximum audio frequency that can cause the deviation is

a. 4 kHz.
b. 8 kHz.
c. 15 kHz.
d. 20 kHz.

10. Assume that one of your unit's FM transmitters has a 60-MHz carrier frequency arid its frequency deviation must not exceed 40 kHz. If the percentage of modulation is 80 percent, the deviation is

a. 20 kHz.
b. 32 kHz.
c. 40 kHz.
d. 64 kHz.

11. Assume that a military FM transmitter is operating on a channel that extends from 92.1 MHz to 92.2 MHz. Since military communication standards require a 10-kHz guard hand on each end of the channel, the maximum permissible frequency deviation is

a. 20 kHz.
b. 40 kHz.
c. 60 kHz.
d. 80 kHz.

12. An FM transmitter being modulated by a 10-kHz audio tone signal has an MI of 5. If the MI is increased to 15 by increasing the amplitude of the modulating signal, the effective bandwidth will be increased from

a. 40 kHz to 70 kHz.

b. 40 kHz to 160 kHz.

c. 160 kHz to 380 kHz.

d. 160 kHz to 480 kHz.

13. Assume that a commercial broadcast FM transmitter has a carrier frequency (fc) of 97.400 MHz and is modulated with a 3-kHz signal. A spectrum analysis indicates that the highest significant frequency (fs) in the output is 97.475 MHz. The MI of this transmitter is

a. 0.5.
b. 7.5.
c. 20.0.
d. 25.0

14. An FM transmitter has a deviation of 20 kHz, and a modulation index of 20 when modulated by a 1-kHz rectangular wave. If the transmitter is to have the same bandwidth when modulated by a 1-kHz triangular wave, it must have a deviation of 3 kHz and a modulation index of

a. 0.3.

b. 1.5.

c. 3.0.

d. 20.

15. When a given FM transmitter is modulated by a 1.5-kHz sine wave, the effective bandwidth is 24 kHz and the frequency deviation is 7.5 kHz. If the transmitter is modulated by a rectangular pulse having a pulse repetition rate of 1.5 kHz and the frequency deviation is held at 7.5 kHz, the modulation index remains at 5 and the effective bandwidth will be

a. 24 kHz.

b. 48 kHz.

c. 60 kHz.

d. 75 kHz.

16. Assume that the preemphasis network in A of figure 1-2 is altered to have a time constant of 75 microseconds. The deemphasis circuit to be used in the receiver is shown in B of figure 1-2. The value of capacitance used in this circuit should be

a. 0.0015 microfarad.

b. 0.0020 microfarad.

c. 0.0100 microfarad.

d. 0.0375 microfarad.

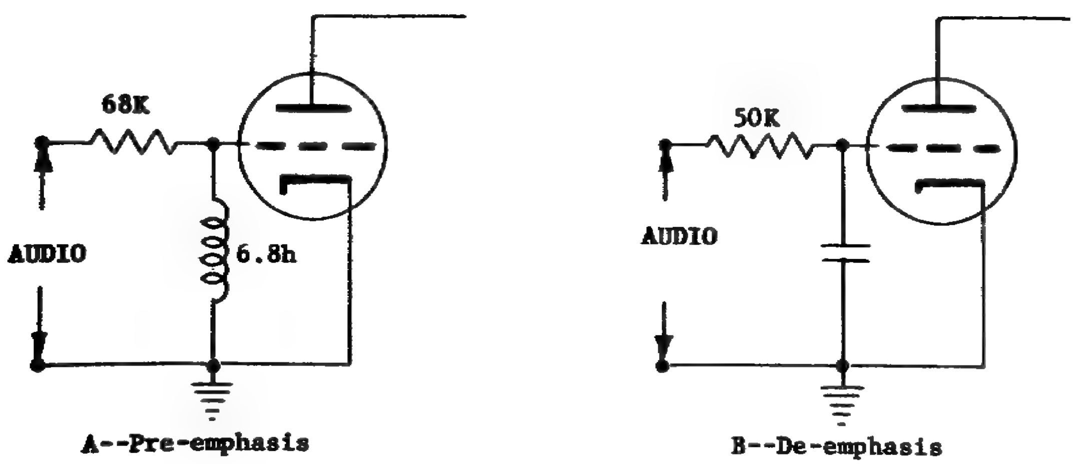


Figure 1-2. Preemphasis and deemphasis circuits.

17. Assume that when a 3-kHz signal is applied to the circuit shown in A, figure 32 of TM 11-668, the signal between the grid and cathode is 4 volts. If the frequency of the input signal is increased to 15 kHz, the signal voltage between the grid and cathode should be

a. 4 volts.

b. 8 volts.

c. 16 volts.

d. 48 volts.

18. The time constant of a preemphasis network is chosen to provide selective amplification of certain frequency bands. The time constant of the preemphasis network shown in figure 1-2 is approximately

a. 45 microseconds.

b. 75 microseconds.

c. 100 microseconds.

d. 150 microseconds.

19. The characteristic of FM that makes it more useful in military applications than AM is its

a. interference-free communication.

b. efficiency at high frequencies.

c. high-fidelity reception.

d. wide acceptance band.

20. Assume that the amplitude of a desired signal at the input of an FM receiver is 2 microvolts, and that other signals are present with frequencies lying within the receiver acceptance band. If the receiver output is to be free of interference noise, the amplitude of the undesired signals must have a maximum limit of approximately

a. 0.002 microvolt.

b. 0.02 microvolt.

c. 1 microvolt.

d. 4 microvolts.

CHECK YOUR ANSWERS WITH LESSON 1 SOLUTION SHEET PAGE 47, 48 and 49.

LESSON 2

METHODS OF PRODUCING FM

SCOPE....................................Study of the methods used to produce direct and indirect FM; circuits used as modulators in direct and indirect FM systems.

CREDIT HOURS.............................2

TEXT ASSIGNMENT..........................TM 11-668, para 18-36;
Attached Memorandum, para 2-1 thru 2-4

MATERIALS REQUIRED.......................None

SUGGESTIONS..............................Read the assignment in TM 11-668 before you read the attached memorandum.

LESSON OBJECTIVES

When you have completed this lesson, you will be able to:

1. Determine the frequency deviation produced by modulated oscillators.

2. Determine the amplification factor, load impedance, and injected reactances for reactance tube modulator stages.

3. Identify and know the operation of the various modulator circuits.

4. Use the block and schematic diagrams of an FM transmitter to determine whether direct or indirect FM is being used.

ATTACHED MEMORANDUM

2-1. ESTABLISHING FREQUENCY MODULATION

a. General. The input and output waveforms of an FM oscillator are shown in figure 2-1. In an FM transmitter, the modulation is accomplished at the oscillator stage. A transistor oscillator can be modulated in the same manner as an electron-tube oscillator or by varying the oscillator gain at the modulating rate. The same amplitude-modulated oscillator used in an AM transmitter can be used in an FM transmitter. The unwanted amplitude changes can be removed by a limiter stage before the carrier signal is increased in frequency and magnitude.

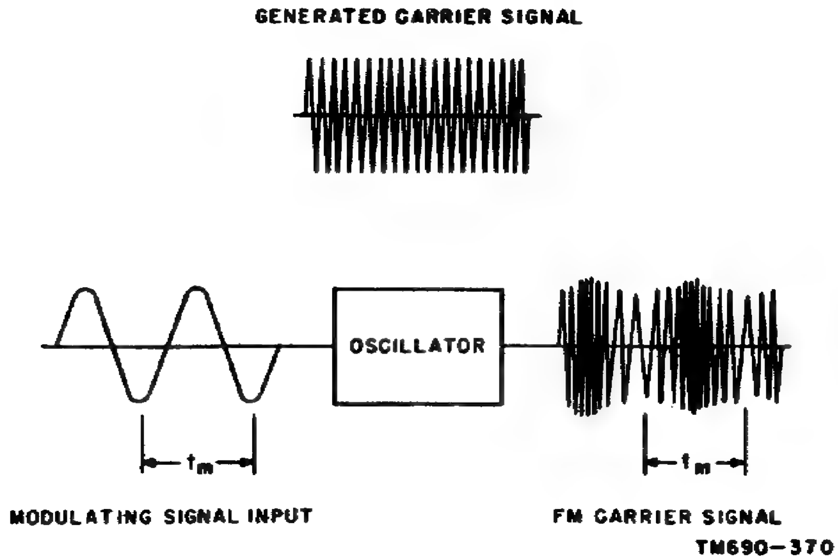


Figure 2-1. FM oscillator, block diagram.

b. FM Oscillator.

(1) A typical FM oscillator stage is shown in figure 2-2. In this application, the modulation is established by reactance modulation. The modulating signal, coupled through transformer T2, varies the emitter-base bias of reactance modulator Q2. Since the bias is increasing and decreasing at the modulation rate, the collector voltage also increases and decreases at the modulating rate. As the collector voltage increases, output capacitance C_{CE} decreases, and as the collector voltage decreases, output capacitance C_{CE} increases. When output capacitance C_{CE} increases, the resonant frequency of the oscillator Q1 tank circuit (capacitor C1 and winding 1-3 of transformer T1) increases. When output capacitance C_{CE} increases, the resonant frequency of the oscillator tank circuit decreases. The resonant frequency of the oscillator tank circuit is therefore increasing and decreasing at the modulating rate, as does the frequency of the signal generated by the oscillator. The output of the oscillator is an FM carrier signal.

(2) Transistor Q1 provides the oscillator signal. Capacitor C1 and winding 1-3 of transformer T1 form a parallel resonant circuit for the oscillator frequency. Winding 4-5 of transformer T1 provides the required feedback, and winding 6-7 couples the oscillator signal to the following stage. Transformer T2 couples the modulating signal to reactance modulator Q2. The reactance of output capacitance C_{CE} across winding 2-3 of transformer T1 varies the resonant frequency of the oscillator tank circuit (capacitor C1 and winding 1-3 of transformer T1).

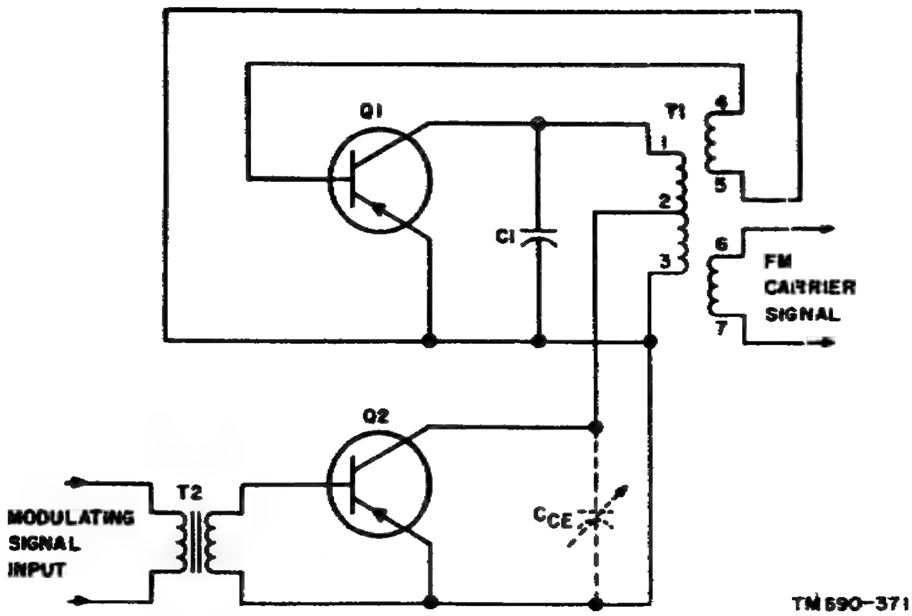


Figure 2-2. Oscillator circuit, frequency-modulated by a reactance modulator.

2-2. USE OF FM OSCILLATOR

An FM oscillator establishes the fundamental frequency-modulated carrier signal necessary for transmission. The requirements for producing an FM carrier using transistors are the same as those for electron tubes. A block diagram of a typical FM transmitter with waveforms is shown in figure 2-3. The audio signal (modulating signal) is applied to the oscillator stage. The output of the oscillator stage is an AM and FM carrier signal. The limiter stage removes the AM and its output is an FM carrier signal with constant amplitude. The multiplier stage increases the frequency to the desired transmitting frequency. The power amplifier stage increases the magnitude of the FM carrier signal sufficiently to drive the antenna.

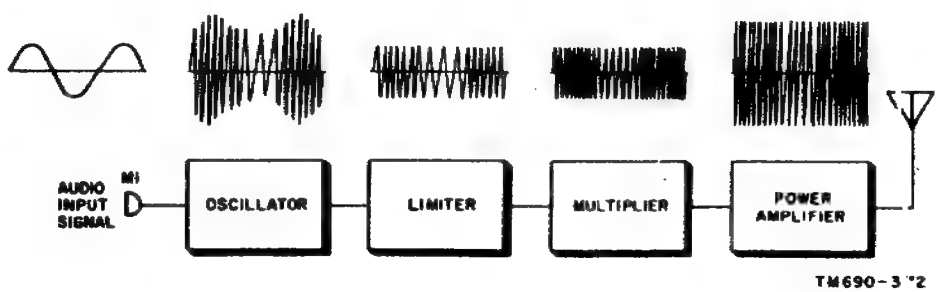


Figure 2-3. FM transmitter, block diagram showing waveforms.

2-3. MILLER-EFFECT MODULATOR

a. Circuit Description. The modulator shown in figure 2-4 employs the Miller effect to vary the equivalent capacitance across a tuned circuit. In the circuit diagram, a complete modulator-oscillator is shown. Transistor Q2 operates as a Colpitts oscillator. Feedback is from the emitter through a capacitively tapped tuned circuit to the base. The operating point is determined by the conventional base voltage divider R6, R7, and emitter bias resistor R8.

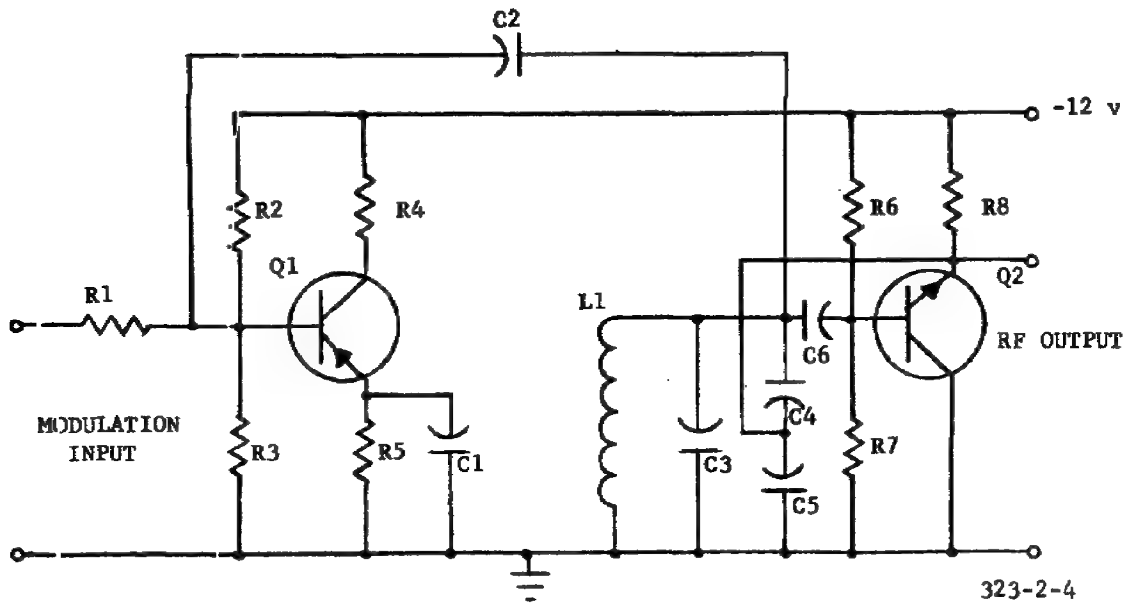


Figure 2-4. Miller-effect modulator.

b. Operation. The modulator, Q1 has an input capacitance at the base that is approximately equal to the emitter junction capacitance plus the voltage gain multiplied by the collector capacitance. If the gain is varied by modulating the operating point of the transistor, the input capacitance of Q1 will change (Miller effect) and frequency modulation occurs.

2-4. REACTANCE MODULATOR ASSEMBLY

a. Speech Amplifier. The transmitter speech amplifier shown in figure 2-5 amplifies the audio-frequency signals from an external microphone or a remote source. The amplified output from this emitter follower is coupled through coupling capacitor C1 and isolating resistor R3 to the oscillator circuit.

b. Oscillator. The output from the speech amplifier modulates the modified Colpitts oscillator to provide an FM output.

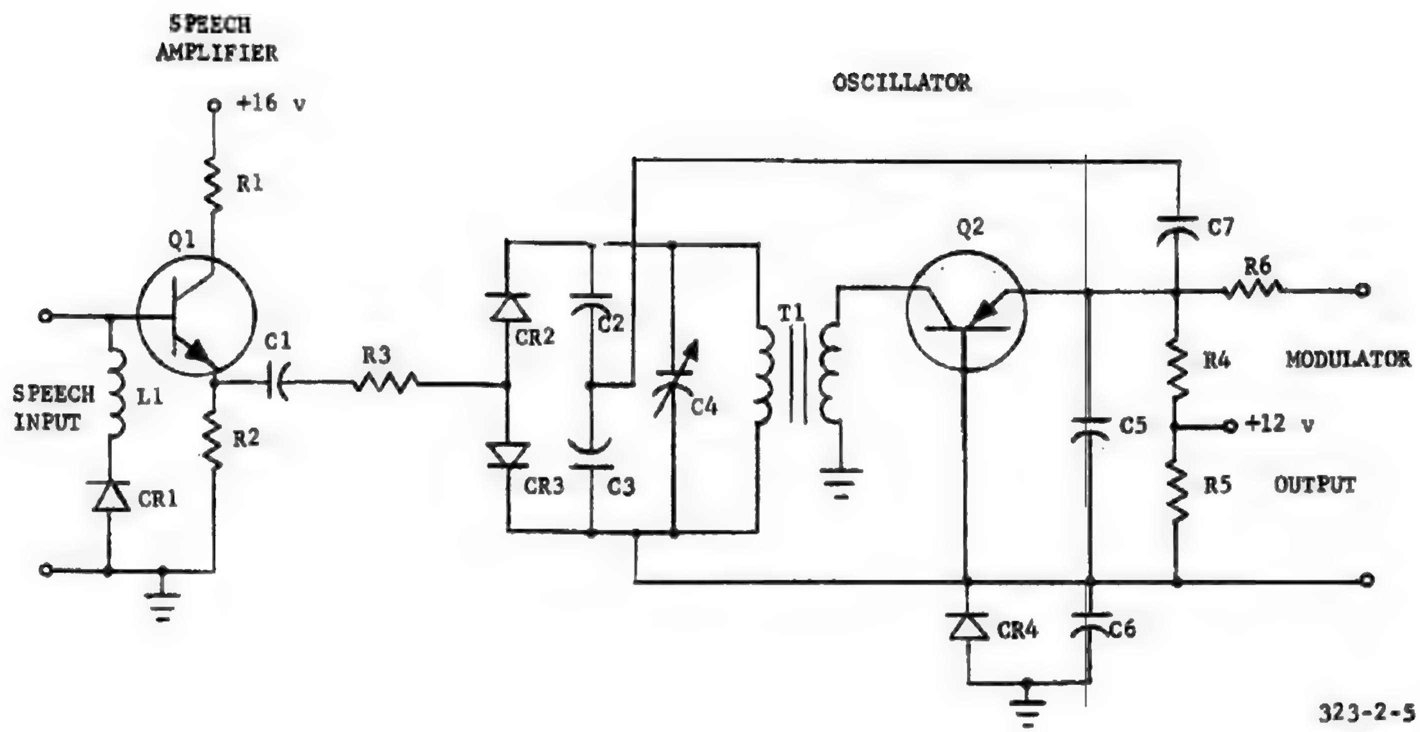


Figure 2-5. Modulator assembly.

(1) The oscillator's tuned circuit consists of transformer T1, trimmer capacitor C4, capacitors C2 and C3, and varactor diodes CR2 and CR3. Capacitors C2 and C3 also serve to match the output impedance of the oscillator to the input impedance of the following stage. Capacitor C7 couples the output back to the emitter to sustain oscillations. Emitter-to-base bias is developed across resistor R4 and C5. Base bias is developed at the junction of R5, zener diode CR4, and C6.

(2) The audio frequency output from the speech amplifier is coupled to the junction of varactor diodes CR2 and CR3. This varying voltage causes the effective capacitance of the diodes to change. These changes in effective capacitance, in accordance with the audio frequency, cause the oscillator frequency to change. Therefore, the output of the oscillator is an FM signal that varies in accordance with the audio frequency from the speech amplifier.

c. Reactance Modulation. The speech amplifier performs the same basis function as the reactance-tube modulator described in TM 11-668. The amplifier changes the incoming amplitude variations of the audio frequency signal into a varying reactance, which is injected into the oscillator's tank circuit. These variations cause changes in the capacitance of CR2 and CR3, which result in capacitive reactance changes in the oscillator's tank circuit.

LESSON EXERCISES

In each of the following exercises, select the ONE answer that BEST completes the statement or answers the question. Indicate your solution by circling the letter opposite the correct answers in the subcourse booklet.

1. Assume that the circuit shown in figure 37 of TM 11-668 is used to produce an FM signal. When the. value of inductance L is 0.9 microhenry (uh) and capacitance C is 0.002 microfarad, the center frequency is 3.75 MHz. If the microphone diaphragm is moved s, that the capacitance is raised to 0.004 microfarad, the frequency deviation will be approximately

a. 0.10 MHz.

b. 1.10 MHz.

c. 2.65 MHz.

d. 3.70 MHz.

2. The Hartley oscillator circuit shown in figure 36 of TM 11-668 is of the type used as the master oscillator in a portable FM transmitter. When the value of coil L is 30 microhenries and the value of capacitor C is 94 picofarads (94×10^{-12} farads), the operating frequency is 3 MHz. If the value of inductance is changed to decrease the frequency to 2.5 MHz, the increase in inductive reactance is approximately

a. 50 ohms.

b. 110 ohms.

c. 300 ohms.

d. 560 ohms.

SITUATION

The reactance-tube modulator used in a short-range, military FM radio transmitter is shown in figure 2-6. Coil L and capacitor C make up the frequency-determining circuit of the Hartley oscillator that provides a center frequency of 22.6 MHz. The characteristics of the modulator tube are as follows:

transconductance (gm) = 5,000 micromhos.

ac plate resistance (r_p) = 0.7 megohm.

Exercises 3 thru 5 are based on this situation.

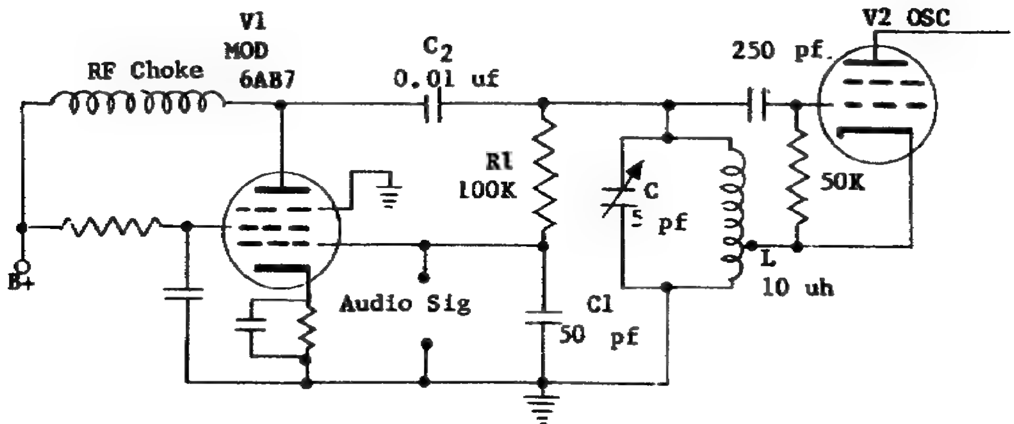


Figure 2-6. Reactance-tube modulator.

3. The electron-tube characteristics described in paragraph 21 of TM 11-668 are computed under static conditions, that is, without a load resistance. The amplification factor of the electron tube used in the modulator circuit is approximately

a. 2,500. c. 5,000.

b. 3,500. d. 7,100.

4. At the center frequency of the oscillator, the impedance across the plate load consisting of R1 and C1 is

a. less than 100K. c. between 125K and 150K.

b. between 100K and 125K. d. greater than 200K.

5. When an audio signal is applied, the reactance-tube modulator adds an inductance to the oscillator tank circuit. The amount of inductance injected by the plate load is

a. 100 millihenry (mh).

b. 50 mh.

c. 10 mh.

d. 1 mh.

6. In the reactance-tube modulator represented by figure 38 of TM 11-668, Z_a is always made large in respect to Z_b because it controls the phase of the current through the plate load. The frequency deviation of a transmitter utilizing this circuit is controlled primarily by the value of

a. impedance in Z_a

b. impedance in Z_b.

c. tank circuit components L and C.

d. transconductance of the reactance tube.

7. Assume that in figure 38 of TM 11-668 the box labeled Z_a represents a 50-picofarad (pf) capacitor, and the box labeled Z_b represents a 1,000-ohm resistor. If the modulator tube has a transconductance of 7,000 micromhos, the reactance component injected into the oscillator tank circuit is equivalent to a

a. 350-uh inductor.

b. 7-uh inductor.

c. 7-pf capacitor.

d. 350-pf capacitor.

8. The circuit shown in figure 43 of TM 11-668 is used to modulate a 40-MHz oscillator. Assume that, with no audio signal applied, this circuit injects an inductance of 7.6 uh into the oscillatory circuit. If L_L = 0.4 mh and R_L = 15K, a tube must be selected that has a transconductance of approximately

a. 1,500 micromhos.

b. 3,500 micromhos.

c. 7,000 micromhos.

d. 9,000 micromhos.

9. Assume that each of the four RC sections of the oscillator in C of figure 48, TM 11-668, has a phase shift of 45° at 4 kHz. When the resistance of the variable resistance leg (modulator tube) is increased, the phase shift for that section is reduced to 30° and the output frequency is lowered. To retain the 180° total phase shift, the phase shift in each of the other RC circuits is now

a. 30°.

b. 45°.

c. 50°.

d. 60°.

10. In producing indirect frequency modulation, the equivalent frequency deviation of the modulated wave is proportional to the modulating frequency. This undesirable effect is eliminated by using a

a. circuit with an output that is inversely proportional to the frequency.

b. phase-shift modulator with a constant impedance network.

c. circuit that functions as a variable resistance.

d. quartz-crystal oscillator with a stable frequency.

11. Which phase modulator would be used in an FM transmitter to provide the greatest initial phase deviation?

a. Link-phase modulator

b. Sonar-phase modulator

c. Reactance-tube phase modulator

d. Nonlinear-coil modulator

12. Assume that the carrier shown in A of figure 56 (TM 11-668) is being modulated by the audio wave shown in B. To obtain a phase-modulated waveform like that shown in F, the output from a nonlinear-coil modulator must be applied to a

a. tuned circuit, RF oscillator, and rectifier.

b. limiter, tuned circuit, and RF oscillator.

c. RF oscillator, rectifier, and limiter.

d. rectifier, limiter, and tuned circuit.

13. Assume that a carrier with a peak amplitude of 10 volts is being phase-modulated by a signal with a peak amplitude of 6 volts. Figure 58 of TM 11-668 indicates the presence of an amplitude component that must be removed by limiting. The peak amplitude of this component is

a. less than 2 volts.

b. between 2 volts and 6 volts.

c. between 6 volts and 10 volts.

d. greater than 10 volts.

14. Frequency modulation of the circuit shown in figure 2-2 is accomplished by reactance modulation. The reactive component that is varied to produce the FM is

a. C_{CE} of Q1.

b. C_{CE} of Q2.

c. T1.

d. C1.

15. Modulation of the Colpitts oscillator shown in figure 2-5 is attained by varying the reactance of components labeled

a. CR2 and CR3.

b. CR2 and T1.

c. C7 and C4.

d. C2 and C3.

16. Some FM transmitters incorporate a limiter stage prior to the multiplier stages. The purpose of this limiter stage is to remove the

a. phase variations from the signal.

b. sideband frequencies from the signal.

c. frequency variations from the signal.

d. amplitude variations from the signal.

17. In the modulator assembly shown in figure 2-5, capacitors C2 and C3 serve a dual purpose--they operate as part of the oscillator tuned circuit and they also

a. develop the oscillator emitter-to-base bias.

b. couple the regenerative feedback to the emitter circuit.

c. emphasize the high frequencies in the modulating signal.

d. match the oscillator impedance to that of the following stage.

18. In the Miller-effect modulator shown in figure 2-4, frequency modulation of the oscillator is produced by injecting the reactive changes from Q1 into the tuned circuit of Q2. The frequency deviation that is produced in Q2 is controlled by the

a. reactance of L1.

b. capacitance of C2.

c. input capacitance of Q1.

d. feedback voltage to the tapped capacitors.

19. All oscillators require that a portion of the output be fed back to the input to sustain oscillations. The component used to provide this feedback in the FM oscillator shown in figure 2-5 is labeled

a. C4.

b. C7.

c. T1.

d. CR4.

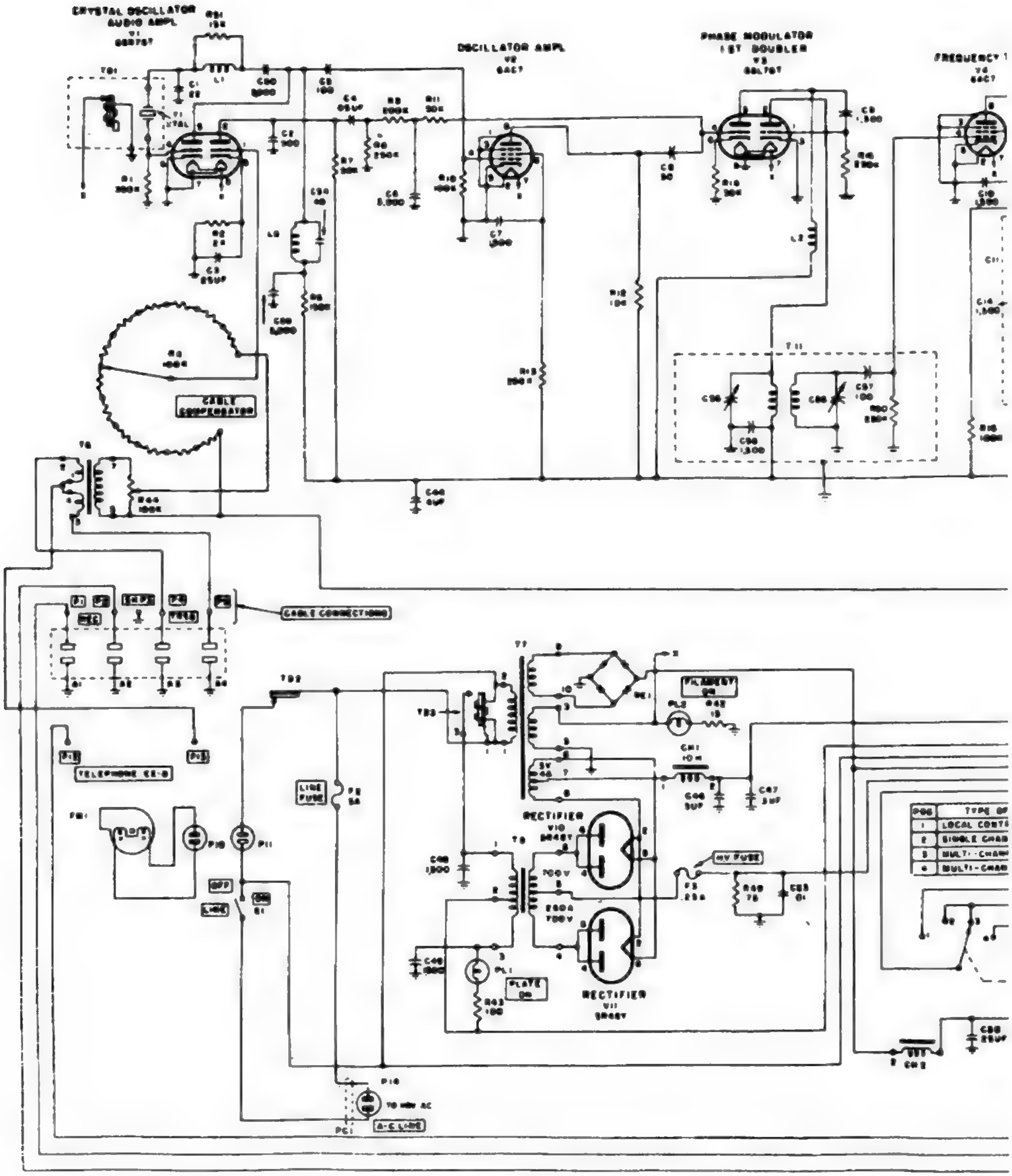


Figure 2-7. FM transmitter.

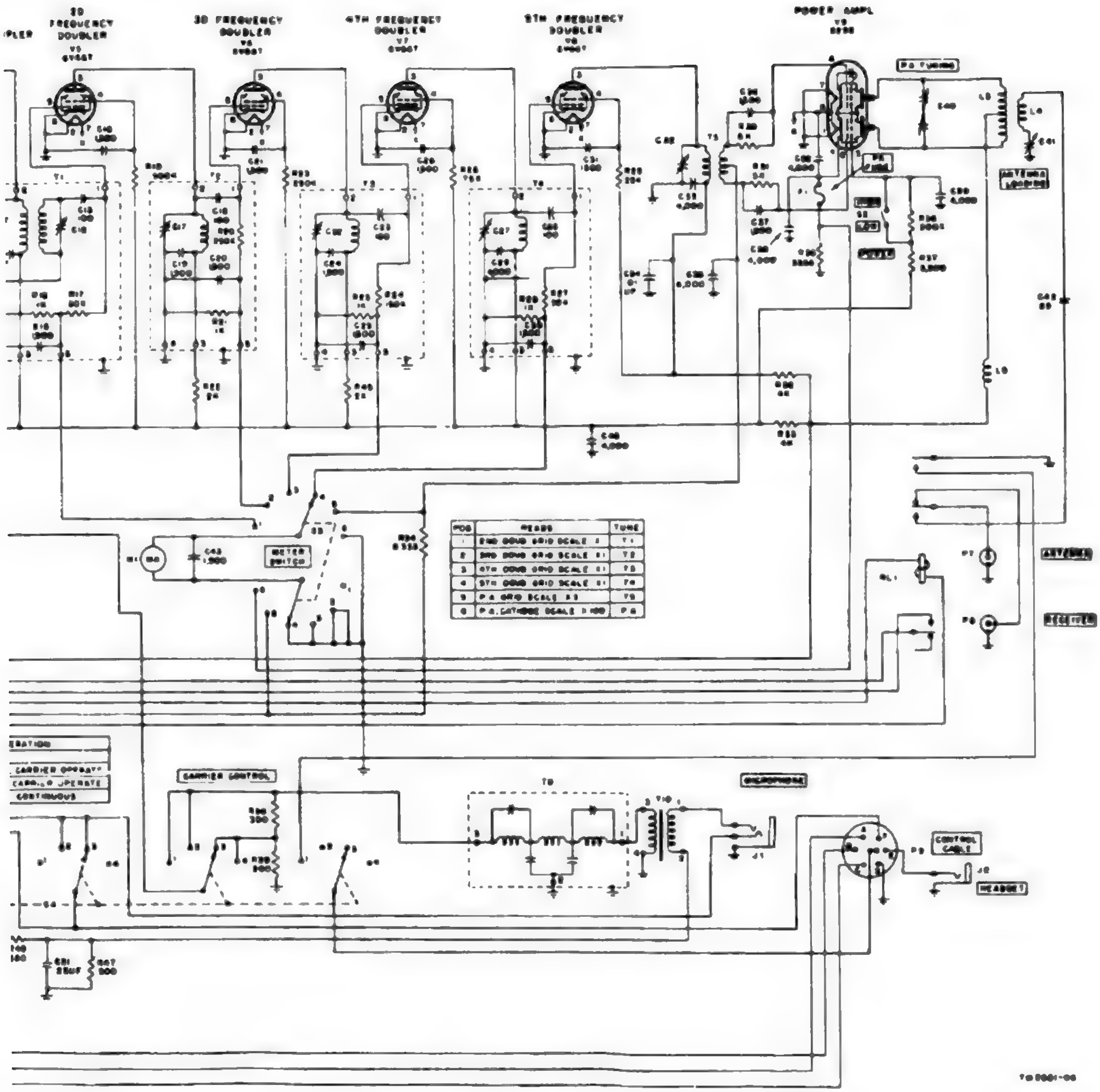


Figure 2-7. FM transmitter. (cont)

20. Assume that the schematic diagram of the FM transmitter shown in figure 2-7 is being used in a classroom to illustrate different types of electron-tube circuits. The type of modulator employed in this transmitter is a

a. reactance-tube modulator.

b. sonar-phase modulator.

c. link-phase modulator.

d. balanced modulator.

CHECK YOUR ANSWERS WITH LESSON 2 SOLUTION SHEET PAGE 49 , 50, 51 and 52.

LESSON 3

FM TRANSMITTER CIRCUITS

SCOPE....................................Principles of operation of frequency multipliers and power amplifiers; impedance matching problems in FM transmitters.

CREDIT HOURS.............................2

TEXT ASSIGNMENT..........................TM 11-668, para 39-41; Attached Memorandum, para 3-1 thru 3-7

MATERIALS REQUIRED.......................None

SUGGESTIONS..............................Read the assignment in TM 11-668 before you read the attached memorandum.

LESSON OBJECTIVES

When you have completed this lesson, you will be able to:

1. Determine the amount of multiplication and the number of multiplier stages needed to produce a given frequency.

2. Explain that frequency multipliers are basically RF amplifiers and how they are tuned to the frequency of the desired harmonic.

3. Describe the input, output, and operating characteristics of the different power amplifier circuits.

4. Calculate the impedances needed for maximum power transfer between stages in the FM transmitter.

ATTACHED MEMORANDUM

3-1. MULTIPLIER THEORY

a. Design Considerations. The frequency multiplier normally consists of a class C amplifier with its output tuned to a multiple of the input frequency. The considerations involved in the design of transistor frequency multipliers are much the same as those in electron-tube class C amplifiers, with one exception--the harmonic content of the collector-current pulse is very sensitive to the phase angle of the collector current flow. The correct collector phase angle must be chosen with respect to the input for the desired frequency ratio.

b. Efficiency. The optimum phase angle, expressed in degrees, is approximately 180 divided by the order of the harmonic; thus, when doubling, 90° should be used. The collector circuit efficiency decreases as the ratios increase and are given approximately 100 divided by the order of the harmonic.

c. Effects on Gain. Unfortunately, the current gain of the transistor used in a multiplier circuit decreases as the frequency and current increase. This causes difficulty in obtaining the desired conduction angles at the higher frequencies. Hence, the collector-current pulses may be broadened because of the random times taken for the holes to diffuse through the base region, and the tops of the pulses may be rounded because of the decrease of the current amplification factor (a_{fe}) with emitter current. These effects are difficult to evaluate; therefore, as a practical matter, an experimental approach is usually made to obtain the best operating conditions. Assume that a transistor is driven, at the base, and the self-bias for the transistor is controlled by a variable time-constant circuit. The phase angle of the collector current can be altered by changing the amplitude of the drive and by changing the time constant. The load impedance can be varied in the collector circuit to effect the optimum transfer of power. If a chain of frequency multipliers is to be used, each stage must be capable of driving the following one. The use of push-pull stages is advantageous with odd-frequency ratios, while the push-push connection is helpful with even ratios. The use of two transistors per stage effectively doubles the power output.

3-2. MULTIPLIER CIRCUIT CHARACTERISTICS

a. Operating Potentials. The frequency multiplier, shown in figure 3-1, is a simple, low-level class C amplifier employing the type of dc stabilization that is normally used with class A stages. The zero signal operating point, 15 volts at 3 milliamperes (ma), is chosen well below the maximum dissipation of the transistor. The collector voltage is made as high as possible without exceeding the breakdown voltage of the transistor on negative collector swings. The transistor used as a multiplier should have good high-frequency response and relatively high collector voltage and dissipation ratings.

b. Circuit Testing. The capabilities of the multiplier circuit are determined by applying a constant input frequency of 1 MHz from a signal generator, which is adjusted to produce the optimum drive amplitude for the particular frequency ratio under test. The load impedance is adjusted by the substitution of various small, high-frequency-type carbon resistors until the maximum power output is obtained. The tuned circuit LC is kept at a medium value of impedance ($X_L = X_C = 300$ ohms) so that a loaded Q of 30 or more can be obtained at the higher harmonics. In the chart below F_{OUT} is the output frequency, R_L is the

F_{OUT} (MHz)	R_L (ohms)	P_{OUT} (mw)	N (%)	Vi (volts)
1	3.3K	37	82	0.4
2	6.8K	21	47	1.0
3	10K	14	31	1.1
4	10K	10	22	1.2
5	10K	6.4	14	1.3
6	10K	3.6	8	1.4
7	10K	2.5	5.5	2.0

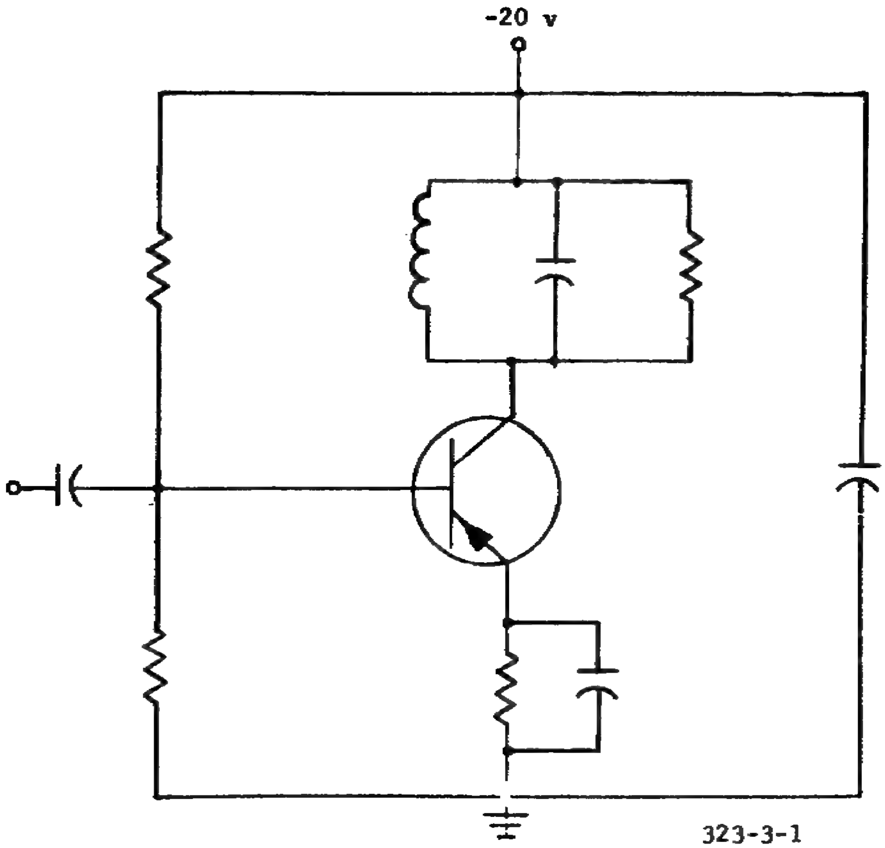


Figure 3-1. Frequency multiplier.

load impedance, P_{OUT} is the power output, N the efficiency, and Vi is the root means square (rms) input voltage. The chart shows that if the frequency ratio (multiplication factor) increases, the power output And efficiency will decrease. You will also note that the input voltage must be increased for each increase in frequency ratio so that a usable output can be obtained. Outputs above the frequency ratio of 5 to 1 are generally not capable of driving the following stage.

c. Limitations. It will be noted that the efficiencies (N) are reasonably close to the expected values (N = 100/n) at the second, third, and fourth harmonics, but become much lower at the higher harmonics. This is probably because of the fact that the minimum collector current pulse duration that could be obtained was 1/8 microsecond, which corresponds to one-half cycle at an output frequency of 4 MHz. At higher output frequencies, this input pulse bridges an excessive portion of the output voltage cycle, and actually extracts energy from the tuned circuit during a portion of the cycle.

3-3. DRIVER AND POWER AMPLIFIER STAGES

Amplifier stages designed primarily to raise the power level of the signal are known as driver and power amplifiers. Driver stages usually develop power

in the order of milliwatts. Power amplifiers develop watts or hundreds of milliwatts of power. This distinction based on power levels is approximate. The power levels of driver and power amplifiers depend on the equipment in which they are used. A driver amplifier, as its name implies, is used to drive a succeeding stage. Thus, the driver stage delivers power to another driver stage or to a power amplifier. Power amplifiers increase the signal power to the necessary level to operate a device such as an antenna.

3-4. SINGLE-ENDED AMPLIFIERS

a. Circuit Arrangements. Circuit arrangements for single-ended driver and power amplifiers do not differ to any marked degree from class A preamplifiers. Drivers and power amplifiers operate at higher collector voltages and currents, however, and are carefully matched for power transfer. Transformer coupling is very useful for matching, and improves efficiency since the dc losses are reduced. Impedance coupling (LC) is sometimes used to obtain a higher efficiency, but it has poor low-frequency response.

b. Efficiency. Efficiency is an important consideration at high power levels, particularly for power amplifiers. As long as we operate class A, efficiency is poor. Both class B and class C operation provide greater efficiency than class A.

3-5. PUSH-PULL AMPLIFIERS

Exact sound reproduction (fidelity) can be realized at higher efficiencies by a push-pull circuit arrangement. Improved efficiency and fidelity can be attained with push-pull amplifiers operating class A. Just as in the case of electron-tube push-pull operation, even-harmonic (nonsymmetrical) distortion is minimized. If the circuit is perfectly symmetrical, such distortion is completely eliminated. Because of this fact, push-pull amplifiers can deliver greater power for an allowable amount of distortion than can two single-ended amplifiers. Moreover, push-pull amplifiers can be operated class B and class AB since even-harmonic distortion can be canceled. Provided there is good circuit symmetry, excellent linearity can be achieved with class AB operation, and fairly good linearity is possible with class B operation.

a. Transformer Type. The class B transformer-coupled amplifier is the simplest type. Note that the NPN transistor circuit shown in figure 3-2 resembles closely the corresponding electron-tube class B push-pull amplifier. Remember that PNP transistors can be used instead, provided the bias polarities are reversed.

(1) Observe in figure 3-2 that there is no forward base-emitter bias; both the base and emitter of Q1 and Q2 are at dc ground potential. Therefore, Q1 and Q2 are cut off; with no-signal input (the static condition) $I_{c1} = 0$ and $I_{c2} = 0$.

(2) A signal applied to the primary of the input transformer produces signals at opposite ends of the center-tapped secondary 180° out of phase, as shown. Therefore, when a negative alternation is applied to the base of Q1, a positive alternation is applied to the base of Q2, and vice versa. Because both Q1 and Q2 have 0 volt base-emitter bias, each will conduct during positive alternations

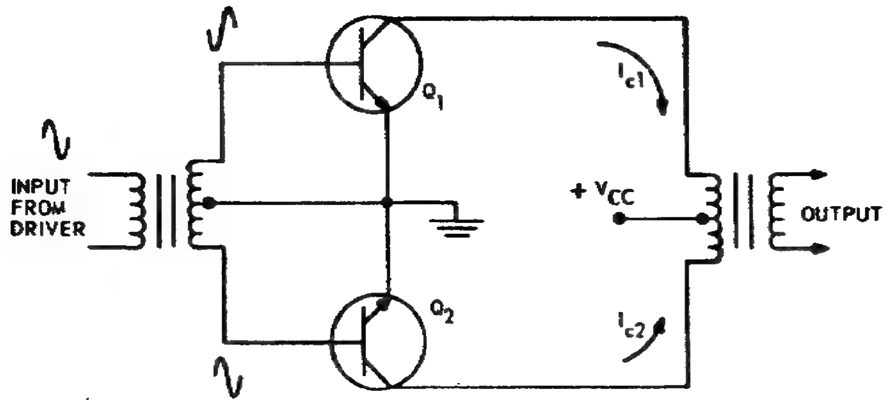


Figure 3-2. Class B push-pull amplifier.

and remain cut off during negative alternations. Since their inputs are 180° out of phase, Q1 is conducting (I_{c1}) while Q2 is cut off, and Q2 is conducting (I_{c2}) while Q1 is cut off. In other words, Q1 and Q2 conduct alternately to supply output current (I_{c1} + I_{c2}) throughout the entire cycle.

(3) This circuit arrangement amounts to having two symmetrically arranged single-ended amplifiers supplying a common load. Sketch A in figure 3-3 is a dynamic curve of a single-ended amplifier; sketch B is the dynamic curve of a push-pull amplifier. Note that the curve in sketch B is merely the combination of the dynamic curves of two single-ended amplifiers. Because of the 180° phase relationship of the input and the circuit arrangement, the positive direction of base currents I_{b1} and I_{b2} are opposite; also, the positive direction of collector currents I_{c1} and I_{c2} are opposite. The input signal wave is projected on the dynamic curve diagram in C of figure 3-3. The output waveshape is a composite of current I_{c1} and I_{c2}. This diagram illustrates clearly the faithfulness of signal reproduction, thus fidelity. Note the distortion that occurs as the signal approaches and passes through zero. This is called crossover distortion and causes odd harmonics of the signal frequency to appear in the output signal.

(4) Crossover distortion can be eliminated by biasing Q1 and Q2 in the forward direction. A simple biasing arrangement is illustrated in figure 3-4. Resistor R is made variable so that we can adjust the bias for class AB operation or class A operation. If properly adjusted for class AB operation, the crossover effect is not evident in the output. A comparison of A in figure 3-5 and B in figure 3-5 shows why this is so. The nonlinearity is effectively canceled and excellent fidelity achieved. If R is adjusted for class A operation, the output waveshape is the resultant of the individual waveshapes as illustrated in C of figure 3-5.

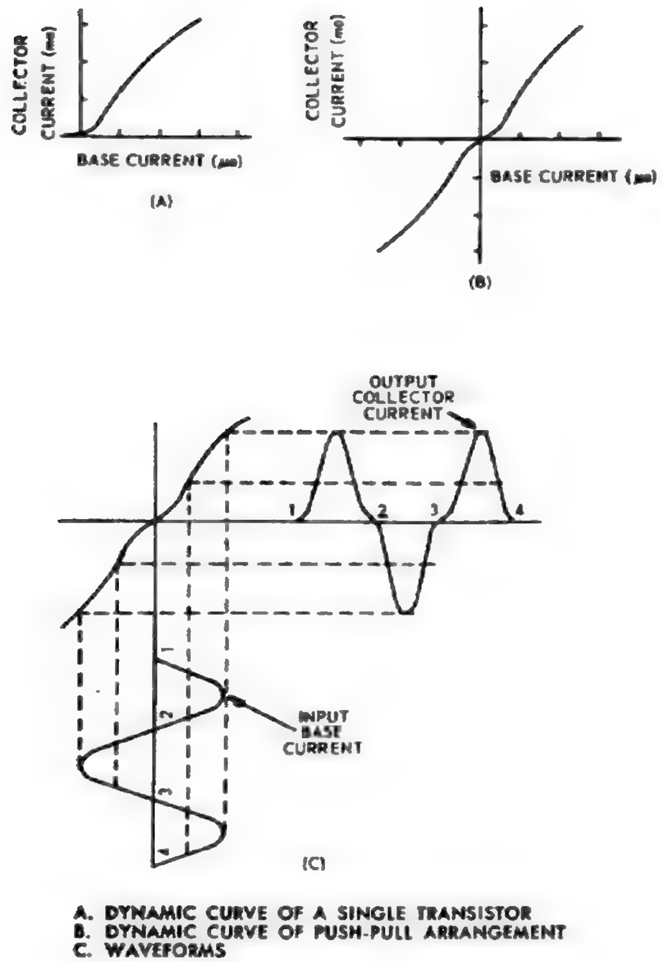


Figure 3-3. Class B push-pull curves.

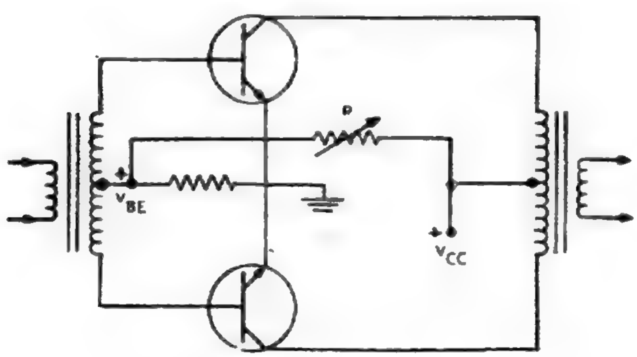


Figure 3-4. Push-pull amplifier that can be biased class AB or A.

(5) Since class AB operation is more efficient than class A operation, it seems that class AB would always be preferred. The output waveshape looks as good for class AB as it does for class A because we have assumed perfectly matched transistors and symmetrical circuitry. Suppose, however, that the transistors are not perfectly matched or that the circuit is not perfectly symmetrical; the dynamic curves for Q1 and Q2 would no longer be identical. For such a condition, the outputs from a class AB and a class A push-pull amplifier are shown in A of figure 3-6 and B of figure 3-6, respectively. It is quite obvious that the output for class AB operation is distorted. This nonsymmetrical distortion is a result of the presence of even harmonics. Thus, we need to realize that class AB operation requires matched transistors and circuit symmetry if high fidelity is desired. Such requirements somewhat offset the higher efficiency that class AB amplifiers offer.

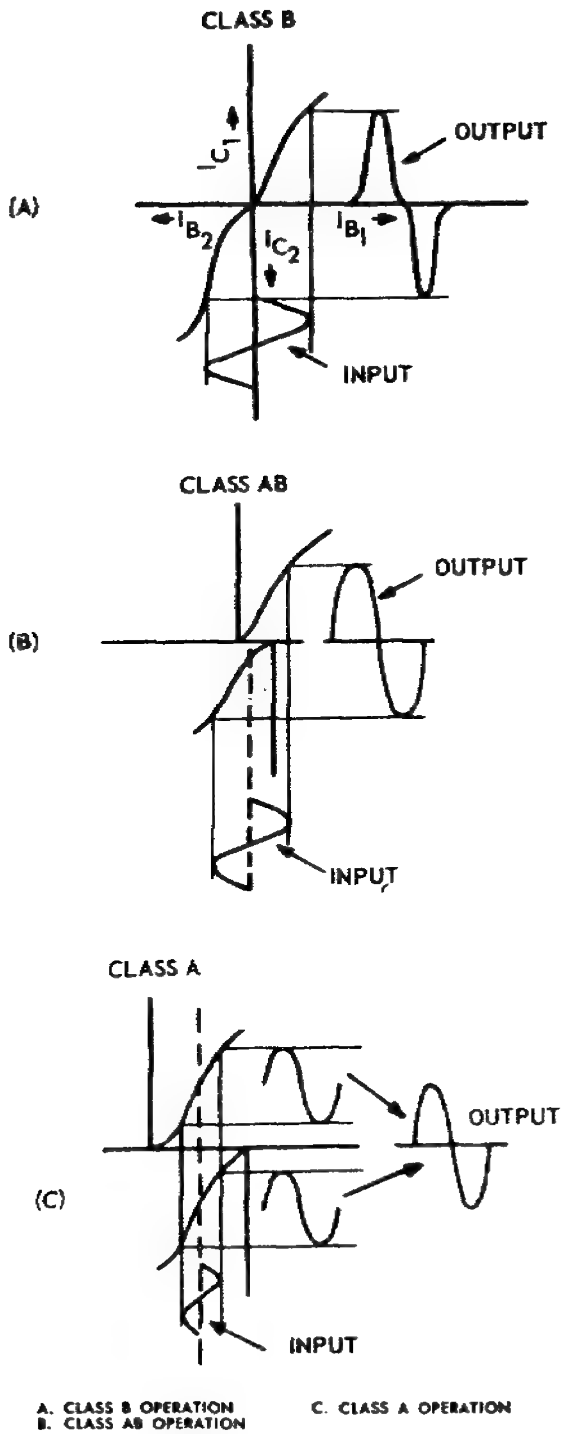


Figure 3-5. Typical output waveforms.

b. Phase Inverter Driver Amplifier. It is not necessary to use a transformer to obtain input signals of opposite phase for a push-pull amplifier. The circuit in figure 3-7 shows how a driver amplifier can be designed to produce output signals that are 180° out of phase. The values of resistors R1 and R2 are chosen to provide the proper base bias for linear class A operation. Resistors R3 and R4 are practically equal in value so that the signal outputs are the same in amplitude. The normal phase inversion from base to collector occurs across R3, whereas across R4 the signal output is in phase with the input.

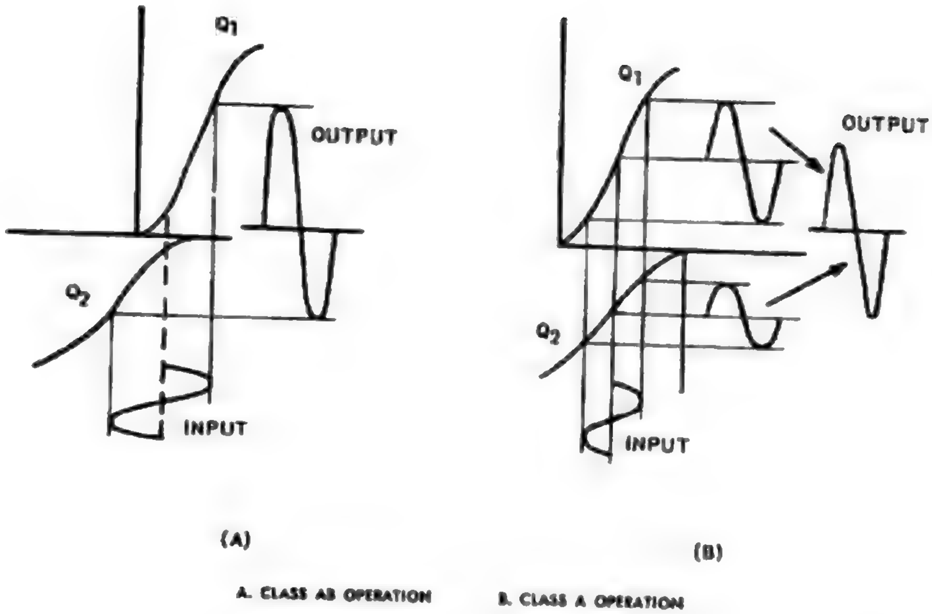


Figure 3-6. Waveforms resulting from nonsymmetrical push-pull circuit.

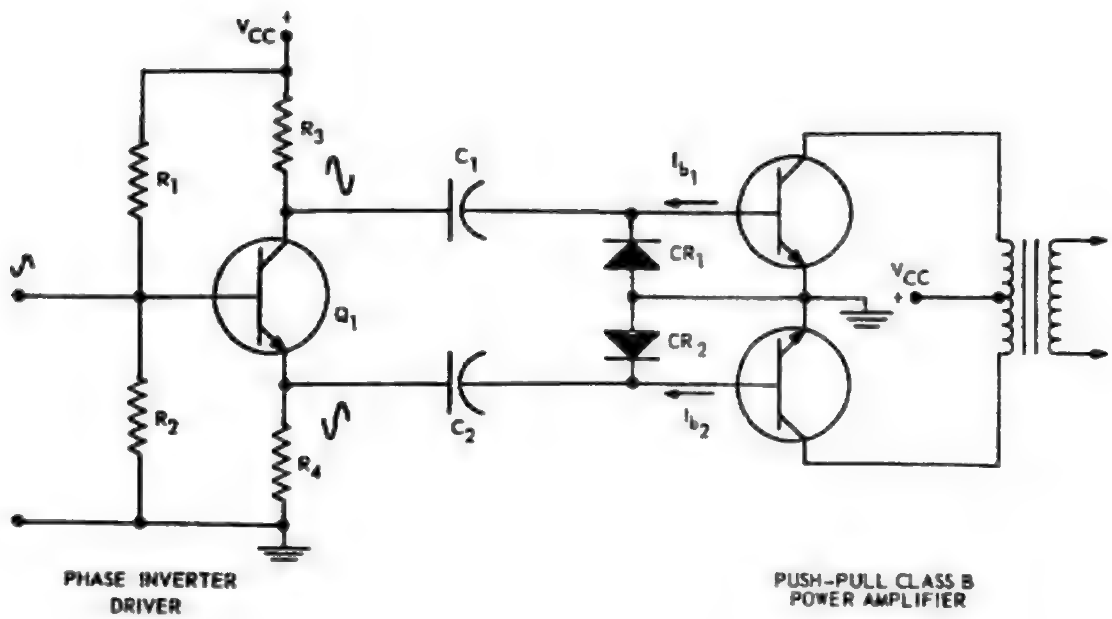


Figure 3-7. Capacitively coupled input to a push-pull amplifier.

(1) Emitter-follower action accounts for the phase of the signal across R4. This action also accounts for the fact that input impedance is desirably higher than ordinary and the output impedance is lower to match the input impedance of the push-pull stage. Another advantage offered is the improved frequency response. Better frequency response can be attributed to two things: degenerative (negative feedback) action and capacitive coupling.

(2) We took care to say capacitive coupling rather than RC coupling, since crystal diodes CR1 and CR2 are used in place of resistors. These are discharge diodes that prevent capacitors C1 and C2 from taking on a charge that would bias the push-pull amplifier well below cutoff. Inasmuch as the transistors draw base current during half a cycle for class B operation, rectified base currents flow into C1 and C2. If CR1 and CR2 were not connected in the circuit as shown, a negative charge would build up to reverse-bias the push-pull transistors. This would correspond to grid-leak biasing of electron tubes. To preclude this unwanted base biasing, CR1 conducts whenever C1 becomes negative with respect to ground, and CR2 conducts whenever C2 becomes negative with respect to ground. Thus, these discharge diodes function to maintain 0 volt base bias.

3-6. COMPLEMENTARY SYMMETRY CIRCUIT

A complementary symmetry circuit arrangement is made possible by the use of an NPN and a PNP transistor. This simple push-pull amplifier circuit shown in figure 3-8, affords the benefits of capacitive coupling, yet does not require a phase inverter nor discharge diodes.

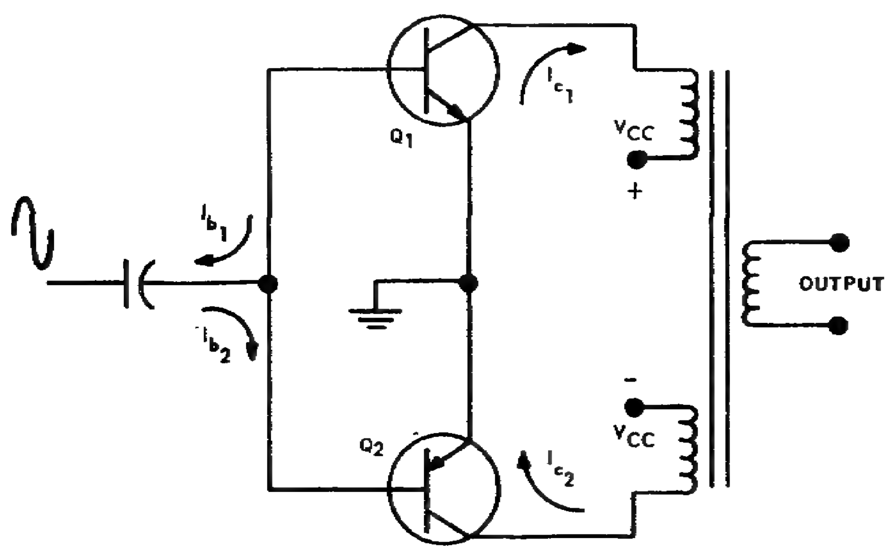


Figure 3-8. Push-pull amplifier using complementary symmetry.

a. Input. Input phase inversion is unnecessary since at zero base bias the NPN transistor (Q1) and PNP transistor (Q2) conduct on alternate half-cycles of the input. When the input signal is positive going, Q1 conducts and Q2 is cut off. When the input signal is negative going, Q2 conducts and Q1 is cut off. Thus, push-pull operation takes place as a result of the complementary action of the transistors.

b. Base Current. Note that I_{b1} and I_{b2} flow in opposite directions. If the transistors are properly matched, these two currents will be equal. Because they are equal and opposite, the resultant charging current flowing into the coupling capacitor is zero. Since no charge will develop to reverse bias the bases, discharge diodes are not needed.

c. Advantage. The basic circuit just discussed illustrates how simplicity and improvement can be attained by using NPN and PNP transistors together. Many varied circuits take advantage of the complementary features of transistors. Such circuits have no electron-tube counterparts.

LESSON EXERCISES

In each of the following exercises, select the ONE answer that BEST completes the statement or answers the question. Indicate your solution by circling the letter opposite the correct answers in the subcourse booklet.

SITUATION

Assume that the FM transmitter shown in figure 2-7 is being modified to increase its efficiency.

Exercises 1 thru 5 are based on this situation.

1. Assume that the crystal in this FM transmitter operates at a frequency of 850 kHz. The center frequency of this transmitter output is

   a. 27.2 MHz.
   b. 54.4 MHz.
   c. 81.6 MHz.
   d. 122.4 MHz.

2. Assume that the transmitter is to be redesigned to have fewer stages and to provide a multiplication of 256. The minimum number of frequency multiplier stages that could be used to obtain this multiplication without excessive loss of output is

   a. three.
   b. four.
   c. five.
   d. six.

3. Assume that the unloaded Q of the power amplifier (V9) tank circuit is 180, and the loaded Q is 9. The transfer efficiency of the tank circuit is approximately

a. 75 percent.

b. 85 percent.

c. 95 percent.

d. 100 percent.

4. After changing V5 to a quadrupler to eliminate the first doubler, it is found that parasitic oscillations occur. These parasitics can be suppressed by inserting a

a. parallel-tuned trap in the screen grid lead.

b. series-tuned circuit in the control grid lead.

c. small high-frequency capacitor between pins 3 and 5 of V5.

d. small resistor in the lead between pin 5 of V5 and pin 1 of transformer T1.

5. Analysis of the transmitter circuit shows that the coupling circuit between the fifth frequency doubler and the power amplifier is of the type that is

a. inductive with link coupling.

b. capacitive with a split-stator tuned circuit.

c. capacitive with the control grid at RF ground.

d. inductive with tuned primary and untuned secondary.

6. Assume that a given portable, direct FM transmitter contains a 20.850-MHz oscillator, a frequency doubler, and a power amplifier. If the frequency of the oscillator swings between the limits of 20.848 MHz and 20.852 MHz while being modulated, the frequency deviation of the transmitter is

a. 2 kHz.

b. 4 kHz.

c. 8 kHz.

d. 10 kHz.

7. Analysis of the i_b - e_c curve for a given electron tube shows that the grid cutoff voltage --8 volts, grid voltage at plate saturation = +2 volts. If this tube is to be used as a doubler in an FM transmitter, its operating bias must be adjusted to approximately

a. -20 volts.

b. -8 volts.

c. -4 volts.

d. +2 volts.

8. Assume that the plate power input to the power amplifier of a low-level AM transmitter is 225 watts and the output power is 55 watts. In an FM transmitter which is also low-level modulated the power amplifier with the same power input is capable of producing a maximum power output of

a. 45 to 90 watts.

b. 90 to 135 watts.

c. 135 to 180 watts.

d. 180 to 225 watts.

9. Since class C amplifiers have tuned grid and plate circuits, they often break into oscillation because of the feedback through interelectrode capacitance. In class C amplifiers used in high-frequency FM transmitters, these oscillations are prevented by

a. using tetrode tubes.

b. decreasing the grid excitation.

c. increasing the Q of the tank circuit.

d. inserting an adjustable capacitor between the plate and grid.

10. A comparison of the single-ended amplifier shown in figure 72 of TM 11-668 and the push-pull amplifier shown in A of figure 74 reveals that the input and output capacitances across the tank circuits of the push-pull circuit are one half those in the single-ended circuit. Since the output voltage of the push-pull circuit is twice that of the single-ended circuit, what is the relationship between the two plate load impedances (Z)?

a. Z's of the two circuits are equal.

b. Z of the push-pull circuit is twice that of the single ended.

c. Z of the single-ended circuit is twice that of the push-pull.

d. Z of the push-pull circuit is four times that of the single ended.

11. A circuit that permits a triode to be operated as a power amplifier at high frequencies without neutralization may be described as having

a. grid-cathode input, plate-cathode output, and grounded cathode.

b. grid-ground input, cathode-ground output, and grounded plate.

c. cathode-ground input, plate-ground output, and grounded grid.

d. grid-cathode input, plate-ground input, and grounded grid.

12. The grid-tank circuit that would be most appropriate for use in a power amplifier located several feet from the driver is shown in TM 11-668, figure 75, sketch

a. C.

b. D.

c. E.

d. G.

13. Assume that the power amplifier in an FM transmitter is driven by a 16-watt signal. The grid draws a current of 18 ma, and the grid bias is -75 volts. For optimum operation, the values of the grid-tank circuit components should be chosen to present an impedance that is

a. less than 10K.

b. between 10K and 100K.

c. between 100K and 1 megohm.

d. greater than 1 megohm.

14. Cross-neutralization of a push-pull amplifier causes an increase in the output capacitance which, in turn, limits the operating frequency. Assume that the push-pull amplifier shown in A of figure 74, TM 11-668, has the values grid-cathode capacitance -6.5 pf, plate-cathode capacitance = 5.5 pf, grid-plate capacitance = 14.5 pf. The INCREASE in output capacitance caused by the neutralizing capacitors is

a. less than 10 pf.

b. between 10 pf and 15 pf.

c. between 15 pf and 20 pf.

d. greater than 20 pf.

15. Sketch A in figure 77 of TM 11-668 shows an ideally tuned power amplifier. Assume that the unloaded Q_O of the tank circuit (L1 -C1) is 200, the operating Q (Q_L) is 10, and the value of X_{L1} is 1,500 ohms. The impedance presented to tie amplifier under load is

a. less than 1K.

b. between 1K and 10K.

c. between 10K and 20K.

d. greater than 20K.

16. Transistors used in multiplier stages should have a good high-frequency response, a high collector dissipation rating, and a high

a. input voltage.

b. input resistance.

c. collector voltage.

d. emitter-to-base bias.

17. Assume that a frequency of 2.5 MHz is applied to a transistorized multiplier stage. If the output of the stage is 10 MHz, what is the approximate efficiency of the circuit?

a. 4 percent

b. 25 percent

c. 50 percent

d. 67 percent

18. Whenever the multiplication factor of the circuit shown in figure 3-1 is changed, several output characteristics also change. A change that is noticed when the multiplication factor is increased is:

a. increase in output efficiency.

b. increase in output power.

c. decrease in output power.

d. decrease in output voltage.

19. The purpose of the discharge diodes in the push-pull amplifier shown in figure 3-7 is to

a. establish a small reverse bias for each transistor.

b. discharge any positive base potential.

c. maintain base bias at 0 volt.

d. rectify the base current.

20. How can you arrange transistor amplifiers to obtain push-pull operation without using a phase inversion stage?

a. Connect two PNP transistors in parallel.

b. Connect two NPN transistors in series.

c. Connect two NPN transistors in a complementary symmetrical arrangement.

d. Connect NPN and PNP transistors in a complementary symmetrical arrangement.

CHECK YOUR ANSWERS WITH LESSON 3 SOLUTION SHEET PAGE 52, 53 and 54.

LESSON 4

FM TRANSMITTERS

SCOPE....................................	Operation of divider and automatic frequency control circuits (AFC); analysis of comparator circuits, to include the double-tuned, phase, and pulse discriminators; circuit analysis of complete FM transmitter.
CREDIT HOURS.............................	2
TEXT ASSIGNMENT..........................	TM 11-668, para 42-51; Attached Memorandum, para 4-1 thru 4-5
MATERIALS REQUIRED.......................	None
SUGGESTIONS..............................	Read the assignment in TM 11-668 and review paragraphs 39-41 before you read the attached memorandum.

LESSON OBJECTIVES

When you have completed this lesson, you will be able to:

1. Analyze the operation of the different types of AFC circuits.

2. State the purpose and explain the operation of frequency-divider circuits.

3. Locate stages in an FM transmitter that perform specified functions.

4. Identify the frequencies and signals that appear at various points in an FM transmitter.

ATTACHED MEMORANDUM

Section I. FREQUENCY DIVIDERS

4-1. SYNCHRONOUS DIVIDER

The frequency dividers of the type under consideration in this section produce a sinusoidal voltage or current at a submultiple of the sinusoidal input frequency. As indicated in the text, waveform devices such as multivibrators or counters can be used, with filtering, to produce a sinusoidal output. However, some nonlinearity must be present to effect division. As a practical matter, a synchronous oscillator is usually involved in the

generation of subharmonics. In one method, a frequency is injected into an oscillator operating at a frequency other than the one injected. It is assumed that a frequency is generated within the oscillator so that a beat frequency can be produced between the two frequencies. Such a simple system is usually termed a locked oscillator and is frequently useful when the frequency ratio (input) is small. A more complicated system can be devised in which the functions of frequency multiplication, beating or phase comparison, and oscillator phase control or locking are carried out in separate portions of the circuit. Although such a device is capable of good performance at large frequency ratios, it is complicated, and will not be considered here.

4-2. LOCKED-OSCILLATOR FREQUENCY DIVIDER

a. Circuit Description. The simple locked-oscillator frequency divider, shown in figure 4-1, is capable of providing reliable frequency division at small frequency ratios. It consists of a grounded base oscillator, with the synchronizing voltage applied between base and ground. Successful operation of this oscillator as a frequency divider requires that harmonics be produced at frequencies near the input frequency. The circuit employs dc stabilization to assure reliable starting.

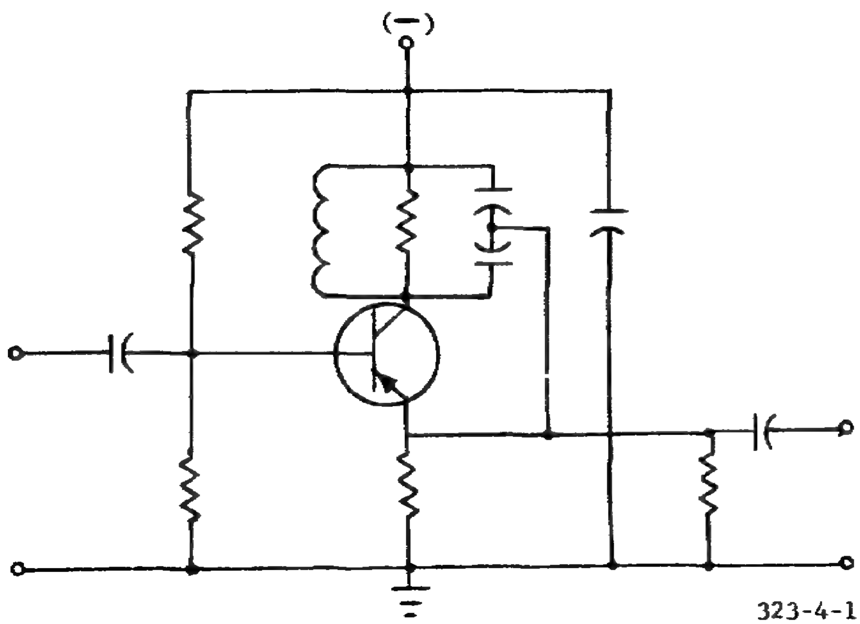


Figure 4-1. Locked-oscillator frequency divider.

b. Operation. The oscillator operates at a frequency that is determined by its frequency-determining network located in the collector circuit. When an input frequency is applied to the base, the oscillator's resonant frequency is altered slightly. In effect, the input frequency locks the oscillator's

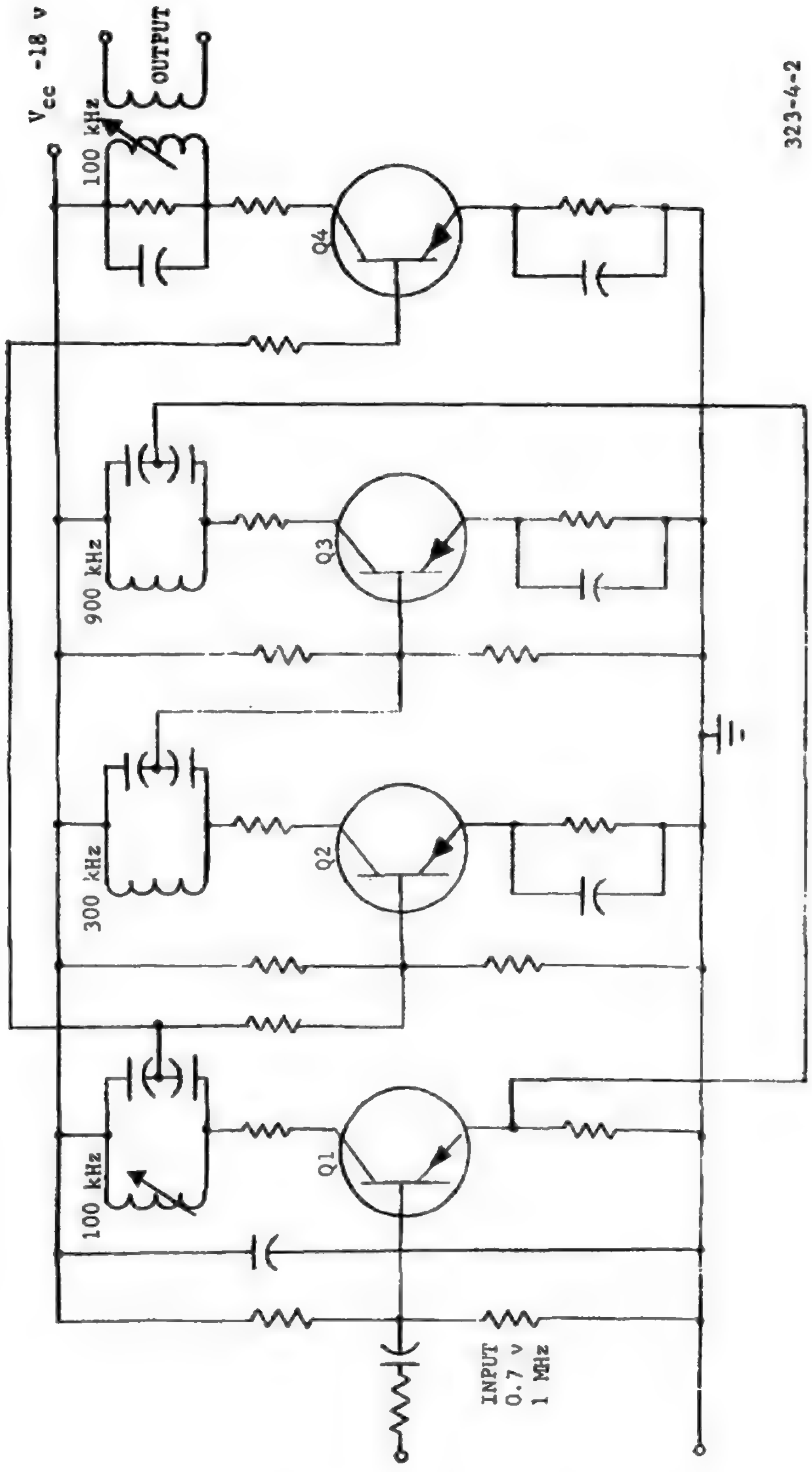


Figure 4-2. Regenerative frequency divider.

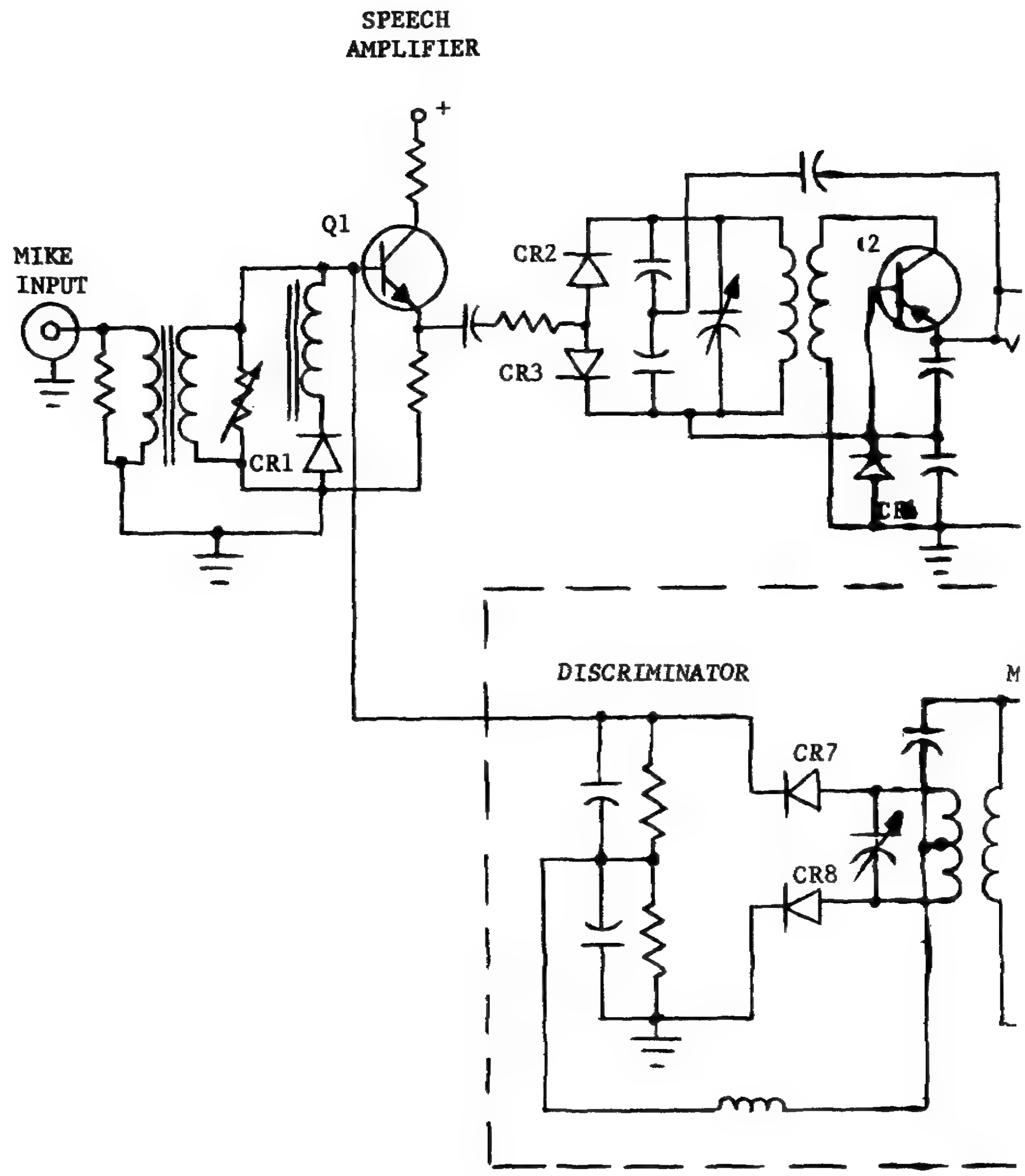


Figure 4-3. Transistorized transmitter.

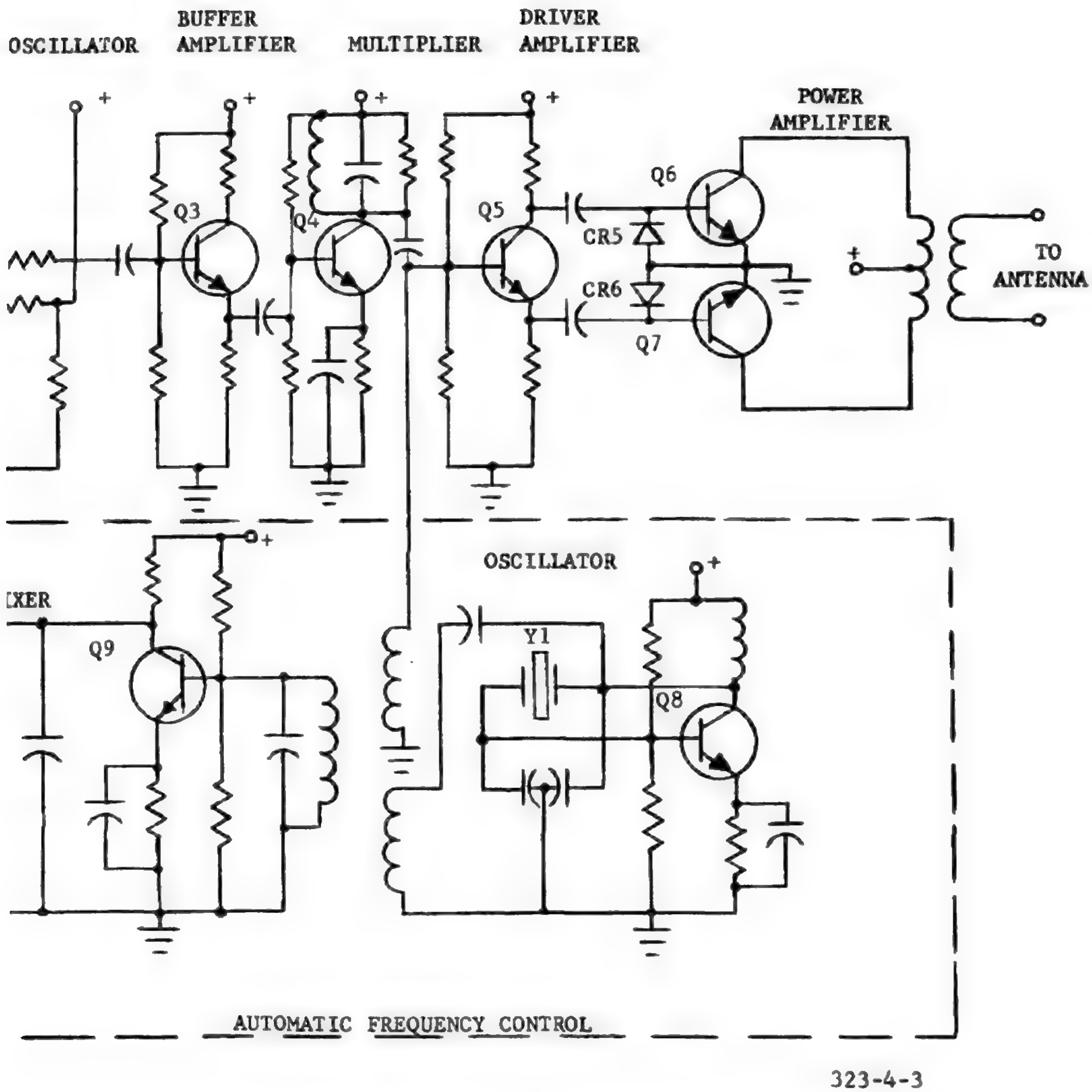


Figure 4-3. Transistorized transmitter. (cont)

frequency at a submultiple of the input frequency. The output frequency is slightly greater than the resonant frequency of the oscillator's tuned circuit.

4-3. REGENERATIVE FREQUENCY-DIVIDER CIRCUIT

a. General. The divider shown in figure 4-2 is driven at a frequency of 1 MHz and produces an output frequency of 100 kHz. The circuit is stable and reliable and, if properly adjusted, will not oscillate in the absence of an input. The phase stability is good and the operating bandwidth is such that excessively close tolerances in the tuned circuit need not be maintained.

b. Circuit Operation. Transistor Q1 is a mixer that receives an input of 0.7 volt rms at an input frequency of 1 MHz from an external source. The output of the mixer is at 100 kHz. The output is processed through two tripler stages to produce an output of 900 kHz. This signal drives the emitter of Q1 to produce a frequency difference of 100 kHz, which is selected by means of a tapped tuned circuit. The resistor connected in series with the collector of the transistor is used to suppress a form of negative resistance oscillation that is encountered with high-frequency junction transistors when bottoming occurs. It also serves the very useful function of limiting the peak collector current to a satisfactory value. Tripler stage Q2 is driven from a capacitive tap on the 100-kHz tuned circuit through a series-isolating resistor which also helps to prevent very high frequency parasitics. The second tripler, Q3, produces the 900-kHz frequency required for the mixer. The output amplifier, Q4, is driven by the mixer output, and has sufficient gain to deliver 20 milliwatts to a 50-ohm load. The resistor, across the tuned circuit of Q4, stabilizes the amplifier and also prevents an excessively high voltage from being developed at this point in absence of a load. The frequency multiplier stages and the output amplifier also contain series-collector resistors for the reasons given above.

c. Alignment. The alignment of the divider is best accomplished by driving the base of each transistor separately at the frequency of the collector tuned circuit. The tuned circuit is then adjusted for maximum response while the input is decreased, if necessary, to avoid limiting. An input of 0.7 volt rms at 1 MHz should then be applied to Q1; the system should oscillate and, as a final step, each tuned circuit should be adjusted to the center of the range over which correct operation is desired. When properly adjusted, the divider should work as the supply is varied from 5 to 40 volts. Increasing the voltage beyond 40 may cause transistor damage, and should not be attempted. The output should be zero in the absence of an input, except for a small amount of noise. The operating bandwidth should be at least ±2 percent at the middle of the supply voltage range.

Section II. TRANSMITTER

4-4. GENERAL

The transmitter illustrated in figure 4-3 is a simplified transistorized version of the one illustrated in figure 100 of TM 11-668. The transmitter comprises circuits that were studied in previous lessons, with the exception of the AFC circuit.

4-5. AFC

The-AFC circuit is used to control the center frequency of oscillator Q2. The circuit compensates for any variations that might be caused by temperature changes, humidity, or vibrations. The circuit contains a crystal oscillator stage, a mixer stage, and a discriminator stage.

a. Crystal Oscillator. Transistor Q8 and its associated components make up a crystal-controlled Colpitts oscillator. The inductance required to resonate with the tapped capacitors is provided by crystal Y1.

b. Mixer. Mixer stage Q9 heterodynes the two signals and produces an out put that is analyzed in the discriminator stage.

c. Discriminator. This phase discriminator produces a dc output voltage that is proportional to the frequency error of the input signal. The polarity of the dc signal depends on whether the input signal is higher or lower than the desired frequency. The dc output voltage is applied to the speech amplifier to control the frequency of oscillator stage Q2.

LESSON EXERCISES

In each of the following exercises, select the ONE answer that BEST completes the statement or answers the question. Indicate your solution by circling the letter opposite the correct answers in the subcourse booklet.

1. Figure 82 of TM 11-668 is a functional block diagram of an FM transmitter and shows an oscillator corrector as part of the AFC system. The electron tube in figure 100 of TM 11-668 that functions as this corrector is

a. V2. c. V4.

b. V3. d. V5.

2. To obtain a zero potential between point A and ground in the circuit shown in figure 84 of TM 11-668, the frequency of the signal in T1 must be equal to the

a. resonant frequency of T2.

b. resonant frequency of T3.

c. difference between the resonant frequencies of T2 and T3.

d. sum of the resonant frequencies of T2 and T3 divided by 2.

3. An electron tube with a single cathode can be used in the double-tuned discriminator shown in figure 84 of TM 11-668. However, the same tube CANNOT be used in the phase discriminator shown in figure 87 because

a. the emission is too low.

b. the circuit becomes unstable.

c. a potential must exist between the two cathodes.

d. high-frequency interference distorts the output.

4. Assume that an AFC system similar to that shown in figure 82 of TM 11-668 uses the basic phase discriminator shown in figure 87 in place of the mixer and discriminator. A positive output voltage of the discriminator indicates that the multiplied input frequency from the master oscillator is

a. equal to the crystal oscillator frequency.

b. less than the crystal oscillator frequency.

c. greater than the crystal oscillator frequency.

d. In phase with that of the crystal oscillator.

5. In the AFC system shown in figure 93 of TM 11-668, a multivibrator is used in the block labeled

a. PHASE SHIFTER.

b. FIRST DIVIDER.

c. MASTER OSCILLATOR.

d. BALANCED MODULATOR.

6. Assume that a signal with a frequency of 440 kHz is to be divided by 4. The oscillator that is to be synchronized by this voltage must have a free-running frequency of

a. less than 110 kHz.

b. between 110 kHz and 220 kHz.

c. between 221 kHz and 440 kHz.

d. greater than 440 kHz.

7. The stages in the AFC system in figure 100 of TM 11-668 include a

a. driver, mixer, and double-tuned discriminator.

b. crystal oscillator, mixer, and phase discriminator.

c. multivibrator, frequency divider, and pulse discriminator.

d. harmonic generator, discriminator, and reactance modulator.

8. Compared with a double-tuned phase discriminator, the advantages of a pulse discriminator include

a. more accurate timing, no frequency-divider circuits, and no stabilizing circuits to tune.

b. no frequency-divider circuits, no stabilizing circuits to tune, and no limiter stages.

c. no stabilizing circuits to tune, no limiter stages, and fewer components.

d. no limiter stages, fewer components, and more accurate timing.

9. Assume that an FM transmitter is equipped with a motor-control AFC system. If the master oscillator drifts to a frequency above the center frequency, the AFC system corrects the situation by

a. modulating the oscillator with a stable frequency.

b. changing the capacitance across the master oscillator tank circuit.

c. injecting an increased amount of inductive reactance across the master oscillator.

d. applying two out-of-place low-frequency signals to the master oscillator.

10. Compared with an indirect FM transmitter, an advantage of a direct FM transmitter is that it generally

a. requires fewer frequency multiplier stages.

b. produces a lower frequency deviation at the oscillator.

c. operates without automatic frequency control.

d. produces a more stable output frequency.

11. An RF amplifier with tuned circuits in the grid and plate will tend to oscillate if the grid circuit receives positive feedback from the plate circuit. The driver stage (V8) of the FM transmitter shown in figure 99 of TM 11-668 is kept from oscillating by the

a. negative feedback voltage that is coupled through C18.

b. transformer coupling between the grid and plate-tank circuits.

c. degenerative feedback voltage developed by C18 and coupled to the input circuit by C17.

d. shield formed between the input and output circuits when the stage is operated in the grounded-grid arrangement.

12. The reason for grounding the center of L7 for RF in figure 100 of TM 11-668 and removing the outputs from the top and bottom of L7 is that it provides a means for developing the

a. bias voltage needed for V11 and V12.

b. out-of-phase input voltages needed for V11 and V12.

c. frequency sample needed for comparison in the AFC circuit.

d. degenerative feedback needed to prevent oscillations in V10.

13. The block diagram in figure 82 of TM 11-668 shows the basic appearance of all AFC circuits. One type of circuit that can be used in the block labeled OSCILLATOR CORRECTOR is the

a. comparator.

b. multiplier.

c. balanced modulator.

d. reactance-tube modulator.

14. Motor-control AFC systems used in FM transmitters require the use of frequency dividers. A typical frequency-divider circuit that uses a two-stage resistance-capacitance-coupled amplifier with the output of the second stage fed back to the input of the first stage is called a

a. synchronized multivibrator.

b. synchronized oscillator.

c. regenerative modulator.

d. trigger circuit.

15. The circuit diagram of an indirect FM transmitter is shown in figure 99 of TM 11-668. The electron tube that is used as a grounded-grid amplifier stage is

a. V3.

b. V4.

c. V7.

d. V9.

16. Basically, there are three frequencies present in the circuit shown in figure 4-1--the input, the output, and the resonant frequencies. What is the relationship between these frequencies?

a. The input frequency is greater than the output frequency but is less than the resonant frequency.

b. The input frequency is less than the output frequency but is greater than the resonant frequency.

c. The input frequency is greater than both the resonant and output frequencies.

d. The input frequency is less than both the resonant and output frequencies.

17. One of the purposes for using the resistors in the collector circuits of Q2 and Q3 in the divider circuit shown in figure 4-2 is to

a. multiply the incoming frequency by 3.

b. heterodyne the input and feedback signals.

c. suppress negative resistance oscillations in the transistors.

d. limit the input frequencies to one third of their original values.

18. Since the output of the divider shown in figure 4-2 is 100 kHz, what is the reason for developing the 900-kHz frequency?

a. To bias mixer stage Q1

b. To lock transistor Q4 at 100 kHz

c. To establish a reference frequency for divider Q2

d. To heterodyne with the input for the development of the 100-kHz output

19. Although the discriminator used in figure 4-3 employs semiconductor diodes, it functions exactly the same as the electron-tube type that is called a

a. balanced modulator discriminator.

b. double-tuned discriminator.

c. pulse discriminator.

d. phase discriminator.

20. If a small dc voltage is applied to the base of Q1 in figure 4-3, either a frequency error or a defective component exists in the transmitter. The frequency error that probably exists is the one generated by

a. oscillator Q2.

b. oscillator QS.

c. audio stage Q1.

d. driver Q5.

CHECK YOUR ANSWERS WITH LESSON 4 SOLUTION SHEET PAGES 54 and 55.

HOLD ALL TEXTS AND MATERIALS FOR USE WITH EXAMINATION AND SIG 324.

LESSON SOLUTIONS

SIGNAL SUBCOURSE 323......................FM Radio Transmitters

LESSON 1.................................Principles of FM

All references are to TM 11-668, unless otherwise indicated.

1. a--para 2, 3

The peak amplitude is the maximum voltage in either the positive or negative direction. Hence, 5 volts.

Figure 1-1 shows 8 cycles in 0.0001 second. Therefore, the number of cycles occurring in one second is 8 x 10,000 = 80,000. The frequency is 80,000 Hz, or 80 kHz

2. a--para 3a

3. b--para 3b(1)

For 100-percent modulation, the modulating voltage must equal the carrier voltage. Therefore, the minimum carrier voltage is 30 volts.

4. b--para 3c(2)

5. b--para 4d(4)

$$\Delta F = \Delta \phi \; f \cos (2\pi f t)$$

$$= \frac{\pi}{6} \times 1{,}200 \times (\pm 1)$$

$$= \pm 628 \text{ Hz}$$

Hence, the carrier varies between the values of 99,372 Hz and 100,628 Hz.

6. a--para 5

7. c--para 7

8. d--para 10c

$$\textbf{Modulation index (MI)} = \frac{\text{maximum frequency deviation}}{\text{maximum frequency of modulating signal}}$$

$$= \frac{40 \text{ kHz}}{5 \text{ kHz}}$$

$$= 8$$

9. b--para 10c

$$\textbf{Modulation index} = \frac{\text{maximum frequency deviation}}{\text{maximum frequency of modulating signal}}$$

$$\textbf{Maximum frequency of modulating signal} = \frac{\text{maximum frequency deviation}}{\text{modulation index}}$$

$$= \frac{40 \text{ kHz}}{5} = 8 \text{ kHz}$$

10. b--para 10d

A deviation of 40 kHz represents 100-percent modulation; therefore, an 80-percent modulation will cause a 32-kHz (40 kHz x 80% = 32 kHz) deviation.

11. b--para 11h, i; Attached Memorandum, para 1-1

Channel width = 92.2 MHz - 92.1 MHz = 0.1 MHz, or 100 kHz

Guard band = 2 x 10 kHz = 20 kHz

Frequency swing = 100 kHz - 20 kHz = 80 kHz

$$\textbf{Frequency deviation} = \frac{\text{frequency swing}}{2} \approx \frac{80 \text{ kHz}}{2} = 40 \text{ kHz}$$

12. c--para 11h(2), table I

Effective bandwidth (MI = 5) = 16 f_A = 16 x 10 kHz = 160 kHz

Effective bandwidth (MI = 15) = 38 f_A = 38 x 10 kHz = 380 kHz

13. c--para 11b(2), table I

fs-fc = 97.475 MHz - 97.400 MHz

= 0.075 MHz, or 75 kHz

Effective bandwidth = 2 x 75 kHz = 150 kHz = 50 f_A

Table I shows that an effective bandwidth of 50 f_A results from the production of 25 sidebands with an MI of 20.

14. c--para 12d(6), (7)

$$\textbf{Modulation index} = \frac{\text{deviation}}{\text{pulse repetition rate}} = \frac{3 \text{ kHz}}{1 \text{ kHz}} = 3.0$$

15. d--para 12d(5), fig. 25, 29

Since the transmitter is pulse-frequency modulated at the same MI of 5, the increase in bandwidth is approximately 400 percent (fig. 29). Hence, the increase in effective bandwidth equals the peak-to-peak deviation times 400 percent (7.5 kHz x 2 x 400% = 60 kHz).

The effective, or total, bandwidth becomes (7.5 kHz x 2) + 60 kHz = 75 kHz.

16. a--para 13d

Time constant = R x C

$$C = \text{Time constant}/R$$

$$C = 75 \times 10^{-6}/50{,}000 = 0.0015 \times 10^{-6}$$

$$C = 0.0015 \text{ microfarad}$$

17. c--para 13c (4), fig. 31

From figure 31 it is seen that a rise from 3 kHz to 15 kHz gives an increase in response of 12 decibels (db)(17 db - 5 db - 12 db). A 12-db increase is equivalent to doubling the voltage twice, or increasing voltage amplitude by 4 to 16 volts.

18. c--para 13c

$$\textbf{Time constant} = \frac{L}{R} = \frac{6.8 \text{ h}}{68{,}000 \text{ ohms}}$$

$$= 0.000100 \text{ second, or } 100 \text{ microseconds}$$

19. a--para 15c

20. c--para 15c

There will be no appreciable interference if the desired signal is twice as large as the interfering signal.

LESSON 2.................................Methods of Producing FM

All references are to TM 11-668, unless otherwise indicated.

1. b--para 18, 19

$$f = \frac{1}{6.28\sqrt{LC}} = \frac{1}{\sqrt{6.28 \quad 0.9 \times 10^{-6} \times 0.004 \times 10^{-6}}}$$

$$= \frac{1}{6.28\sqrt{0.0036 \times 10^{-12}}} = \frac{1}{6.28 \times 0.06 \times 10^{-6}}$$

$$= 2.65 \text{ MHz.}$$

Frequency deviation - 3.75 MHz - 2.65 MHz - 1.10 MHz.

2. b--para 18, 20

At 3 MHz, X_L - $2\pi fL$ - 6.28 x (3 x 10^6) x (30 x 10^{-6}) - 565 ohms.

Since $X_L = X_C$ in a resonant circuit, and the capacitance is unchanged

at 2.5 MHz, $X_L = X_C = \frac{1}{2\pi fC} = \frac{1}{6.28 \times 2.5 \times 10^6 \times 94 \times 10^{-12}} = 678$ **ohms.**

The required increase in inductance reactance is 678 ohms - 565 ohms = 113 ohms, or approximately 110 ohms.

3. b--para 21d

Amplification factor (μ) = gm x r_p

= 5,000 x 10^{-6} x 0.7 x 10^6

= 3,500

4. c--para 22b(3)(as changed), 22e(2)

$$Zab = \frac{1}{gm}\left(1 + \frac{Za}{Zb}\right)$$

$$Zab = \frac{1}{5{,}000 \times 10^{-6}}\left(1 + \frac{R1}{X_{C1}}\right)$$

$$X_{C1} = \frac{1}{2\pi fC1} = \frac{1}{6.28 \times 22.6 \times 10^6 \times 50 \times 10^{-12}} = 141 \text{ ohms}$$

$$Zab = 200\left(1 + \frac{100{,}000}{141}\right)$$

$$Zab = 200\ (1 + 709) = 142{,}000 \text{ ohms, or } 142K$$

5. d--para 22e(2)

$$Li = \frac{R_L\ C_L}{gm} = \frac{100 \times 10^3 \times 50 \times 10^{-12}}{5{,}000 \times 10^{-6}} = 1 \text{ mh}$$

6. d--pare 22e(5)

7. d--para 22d(2)

The arrangement of the plate-load components in figure 38 causes a capacitive reactance to be injected. The Injected capacitance is:

Ci - gm x R x C

Ci - 7,000 x 10-6 x 1,000 x 50 x 10^{-12}

Ci - 350 pf

8. b--para 22e(4)

$$Li = \frac{L_L}{gm\ R_L}$$

$$7.6 \times 10^{-6} = \frac{0.4 \times 10^{-3}}{gm \times 15 \times 10^{3}}$$

$$gm = \frac{0.4 \times 10^{-3}}{15 \times 10^{3} \times 7.6 \times 10^{-6}} = \frac{0.4}{114}$$

gm = 0.0035 mhos, or 3,500 micromhos

9. c--para 26

The total phase shift is 180° at the oscillating frequency. Hence, when the increased variable resistance reduces the phase shift for one section, the remaining sections increase their phase shift when the frequency adjusts to a lower value.

10. a--para 30

11. b--para 31f, 32c, 33c, 34a

12. d--para 34g

13. a--para 35d, fig. 57, 58

From A of figure 58, TM 11-668, the amplitude of this AM component is equal to vector AD minus vector AB.

$$AD - AB = \sqrt{10^2 + 6^2} - 10$$

$$= \sqrt{136} - 10$$

$$= 11.7 - 10$$

$$= 1.7 \text{ volts}$$

14. b--Attached Memorandum, para 2-1b

15. a--Attached Memorandum, para 2-4b(2)

16. d--Attached Memorandum, para 2-2

17. d--Attached Memorandum, para 2-4b(1)

18. c--Attached Memorandum, para 2-3b

19. b--Attached Memorandum, para 2-4b(1)

20. c--para 31; fig. 51

LESSON 3................................FM Transmitter Circuits

All references are to TM 11-668, unless otherwise indicated.

1. c--para 40c(5)

$$\text{Output frequency} = 2 \times 3 \times 2 \times 2 \times 2 \times 2 \times \text{crystal frequency}$$
$$= 96 \times 850 \text{ kHz}$$
$$= 81.6 \text{ MHz}$$

2. b--para 40c(5), (6)

The maximum multiplication by a single stage is five, but five is ruled out because it is not a proper factor of 256. The multiplication factor of 256 can be attained by using four quadrupler stages (4 x 4 x 4 x 4 = 256).

3. c--para 41h(4)

$$\text{Transfer efficiency} = \frac{Q_O - Q_L}{Q_O} \times 100$$
$$= \frac{180 - 9}{180} \times 100 = \frac{17,100}{180} = 95 \text{ percent}$$

4. d--para 41k(4)

5. d--para 41g

6. b--para 40b

Upper frequency limit = 2 x 20.852 MHz - 41.704 MHz

Center frequency = 2 x 20.850 MHz = 41.700 MHz

Frequency deviation = 41.704 MHz = 41.700 MHz

= 0.004 MHz, or 4 kHz

7. a--para 40c(2), (3), (4)

8. c--para 41b(4)

The power amplifier of a low-level AM transmitter must provide linear amplification which in this case is about 25 percent efficient ($\frac{\text{55-watt output}}{\text{225-watt input}}$), whereas the power amplifier of an FM transmitter may be operated at the more efficient class C. Since the plate efficiency of a class C amplifier is from 60 percent to 80 percent,

Power out_1 = 0.60 x 225 watts - 135 watts

Power out_2 - 0.80 x 225 watts = 180 watts

9. a--para 41d

10. d--para 41f(3)

11. c--para 41e

12. d--para 41g(10)

13. b--para 41g(2)

$$\text{Grid-tank circuit impedance} = \frac{\text{driving power}}{(i_g)^2}$$

$$= \frac{16 \text{ watts}}{(0.018)^2}$$

$$= \frac{16}{0.000324}$$

$$= 50K$$

14. d--para 41f(1), (2)

The output capacitance of an unneutralized push-pull amplifier is one-half the output capacitance of one tube. The output capacitance of a neutralized push-pull amplifier is one-half the output capacitance of one tube plus twice the grid-plate capacitance. Hence,

increase in capacitance = 2 x (grid-plate capacitance)

= 2 (14.5 pf)

= 29 pf.

15. c--para 41h(4)

Load impedance = Q_L x X_{L1}

= 10 x 1,500

= 15,000 ohms, or 15K

16. c--Attached Memorandum, para 3-2a

17. b--Attached Memorandum, para 3-2c

To produce an output of 10 MHz with a 2.5-MHz input, the collector circuit must be tuned to the fourth harmonic $(\frac{10 \text{ MHz}}{2.5 \text{ MHz}} = 4)$. Therefore, the approximate efficiency $N = \frac{100}{n} = \frac{100}{4} = 25$ **percent**

18. c--Attached Memorandum, para 3-2

19. c--Attached Memorandum, para 3-5b(2)

20. d--Attached Memorandum, para 3-6a

LESSON 4.................................FM Transmitters

All references are to TM 11-668, unless otherwise indicated.

1. a--para 42b

2. d--para 43b

3. c--para 43d(1)

Using a single cathode would short the output load circuit. When the frequency varies from the center frequency in a discriminator, a potential difference must appear between the two cathodes to obtain an output voltage that is proportional to the frequency change.

4. c--para 43d(2)

5. b--para 45b(1)

6. a--para 45c

The output is 1/4 x 440 kHz, or 110 kHz. However, the free-running frequency must be less than the output for greatest stability.

7. b--para 47g

8. a--para 43f(2), (11), (12)

9. b--para 44c

10. a--para 47a

11. d--para 41e, 46g, h

12. b--para 47h

13. d--para 43c(1)

14. a--para 45b(1)

15. d--para 46g

16. c--Attached Memorandum, para 4-2b

17. c--Attached Memorandum, para 4-3b

18. d--Attached Memorandum, para 4-3b

19. d--para 43d, 47g; Attached Memorandum, para 4-5c

20. a--Attached Memorandum, para 4-5

*May 1971

CORRECTIONS TO TM 11-668

Page 9, para 4d(4), lines 30 and 31. Change the formula to:

$$\underline{\Delta F} = \frac{\pi}{6} \times 1{,}000 \times (+1)$$

$$\underline{\Delta F} = +523 \text{ cps (approximately).}$$

Page 20, para 11g(2), line 5. Change "25" to: 15.

Page 29, para 13d, make the following changes:

Line 16. Change "750,000 to: 75,000.

Line 18. Change "750,000" and ".001" to: 75,000 and $.001 \times 10^{-6}$, respectively.

Page 36, para 18c, line 1. Change "shunt-fed" to: series-fed.

Figure 36, caption. Change "Shunt fed" to: Series-fed.

Page 37, para 20, first formula. Change to:

$$X_C = \frac{1}{2\pi fC} \text{ ohms.}$$

Page 40, para 22b(3), formula at top of page. Change to:

$$Z_{ab} = \frac{1}{g_m} \times \left(\frac{Z_a + Z_b}{Z_b}\right)$$

$$Z_{ab} = \frac{1}{g_m} \times \left(1 + \frac{Z_a}{Z_b}\right) \qquad g_m = \text{mhos}$$

$$Z_{ab} = \frac{1}{g_m} + \frac{Z_a}{g_m Z_b}$$

Page 94, para 43d(3). Delete line 7, and substitute: less; therefore, a positive voltage.

Page 118, para 59a, lines 17-25. Change to: The over-all output is the quadrature sum of the signal and the noise voltages, multiplied by the stage amplification, or

$$\sqrt{10^2 + 4.6^2} \times 10 = \underline{10.9} \times 10 = 109 \text{ \underline{microvolts}}.$$

*This edition replaces correction sheet dated April 1969.

EXTRACT OF TM-11-668
USED IN SS 323 Ed 6 ONLY

CHAPTER 1
MODULATION SYSTEMS

1. Introduction

a. In the armed services, a reliable radio communication system is of vital importance. The swiftly moving operations of modern armies require a degree of coordination made possible only by radio. Today, the two-way radio is standard equipment in almost all military vehicles, and the walkie-talkie is a common sight in the infantry. Until recently, a-m (*amplitude modulation*) communication was used universally. This system, however, has one great disadvantage: Random noise and other interference can cripple communication beyond the control of the operator. In the a-m receiver, interference has the same effect on the r-f signal as the intelligence being transmitted because they are of the same nature and inseparable.

b. The engines, generators, and other electrical and mechanical systems of military vehicles generate noise that can disable the a-m receiver. To avoid this difficulty a different type of modulation, such as p-m (phase modulation) or f-m (frequency modulation), is used. When the amplitude of the r-f (radio-frequency) signal is held constant and the intelligence transmitted by varying some other characteristic of the r-f signal, some of the disruptive effects of noise can be eliminated.

c. In the last few years, f-m transmitters and receivers have become standard equipment in the Signal Corps, and their use in mobile equipments exceeds that of a-m transmitters and receivers. The widespread use of frequency modulation means that the technician must be prepared to repair a defective f-m unit, aline its tuned circuits, or correct an abnormal condition. To perform these duties, a thorough understanding of frequency modulation is necessary.

2. Carrier Characteristics

The r-f signal used to transmit intelligence from one point to another is called the *carrier*. It consists of an electromagnetic wave having amplitude, frequency, and phase. If the voltage variations of an r-f signal are graphed in respect to time, the result is a waveform such as that in figure 2. This curve of an unmodulated carrier is the same as those plotted for current or power variations, and it can be used to investigate the general properties of carriers. The unmodulated carrier is a sine wave that repeats itself in definite intervals of time. It swings first in the positive and then in the negative direction about the time axis and represents changes in the amplitude of the wave. This action is similar to that of alternating current in a wire, where these swings represent reversals in the direction of current flow. It must be remembered that the plus and minus signs used in the figure represent direction only. The starting point of the curve in figure 2 is chosen arbitrarily. It could have been taken at any other point just as well. Once a starting point is chosen, however, it represents the point from which time is measured. The starting point finds the curve at the top of its positive swing. The curve then swings through 0 to some maximum amplitude in the negative direction, returning through 0 to its original position. The changes in amplitude that take place in this interval of time then are repeated exactly so long as the carrier remains unmodulated. A full set of values occurring in any equal period of time, regardless of the starting point, constitutes one cycle of the carrier. This can be seen in the figure, where two cycles with different starting points are marked off. The number of these cycles that occur in 1 second is called the *frequency* of the wave.

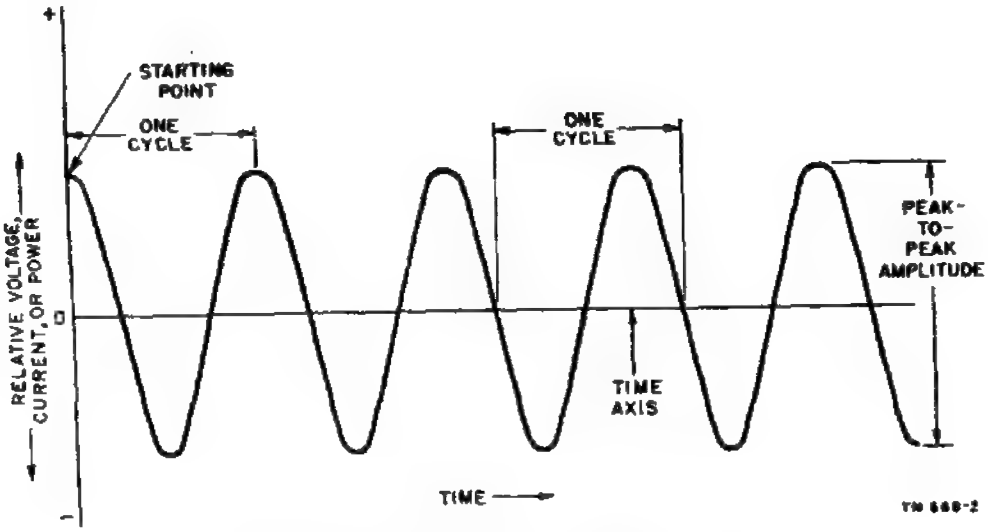


Figure 2. Graph of typical unmodulated carrier.

3. Amplitude Modulation

a. General. The amplitude, phase, or frequency of a carrier can be varied in accordance with the intelligence to be transmitted. The process of varying one of these characteristics is called *modulation.* The three types of modulation, then, are amplitude modulation, phase modulation, and frequency modulation. Other special types, such as pulse modulation, can be considered as subdivisions of these three types. With a sine-wave voltage used to amplitude-modulate the carrier, the instantaneous amplitude of the carrier changes constantly in a sinusoidal manner. The maximum amplitude that the wave reaches in either the positive or the negative direction is termed the *peak amplitude.* The positive and negative peaks are equal and the full swing of the cycle from the positive to the negative peak is called the *peak-to-peak amplitude.* Considering the peak-to-peak amplitude only, it can be said that the amplitude of this wave is constant. This is a general amplitude characteristic of the unmodulated carrier. In amplitude modulation, the peak-to-peak amplitude of the carrier is varied in accordance with the intelligence to be transmitted. For example, the voice picked up by a microphone is converted into an a-f (audio-frequency) electrical signal which controls the peak-to-peak amplitude of the carrier. A single sound at the microphone modulates the carrier, with the result shown in figure 3. The carrier peaks are no longer constant in amplitude because they follow the instantaneous changes in the amplitude of the a-f signal. When the a-f signal swings in the positive direction, the carrier peaks are increased accordingly. When the a-f signal swings in the negative direction, the carrier peaks are decreased. Therefore, the instantaneous amplitude of the a-f modulating signal determines the peak-to-peak amplitude of the modulated carrier.

b. Percentage of Modulation.

(1) In amplitude modulation, it is common

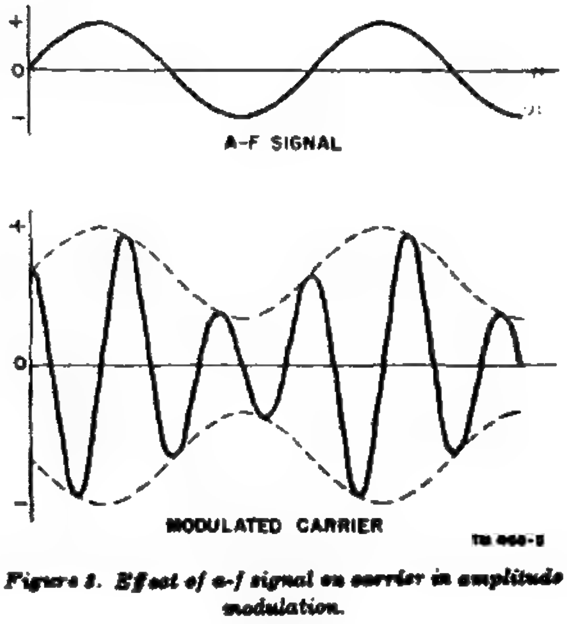


Figure 3. Effect of a-f signal on carrier in amplitude modulation.

practice to express the degree to which a carrier is modulated as a percentage of modulation. When the peak-to-peak amplitude of the modulating signal is equal to the peak-to-peak amplitude of the *unmodulated* carrier, the carrier is said to be 100 percent modulated. In figure 4, the peak-to-peak modulating voltage, E_A, is equal to that of the carrier voltage, E_R, and the peak-to-peak amplitude of the carrier varies from $2E_R$, or $2E_A$, to 0. In other words, the modulating signal swings far enough positive to double the peak-to-peak amplitude of the carrier, and far enough negative to reduce the peak-to-peak amplitude of the carrier to 0.

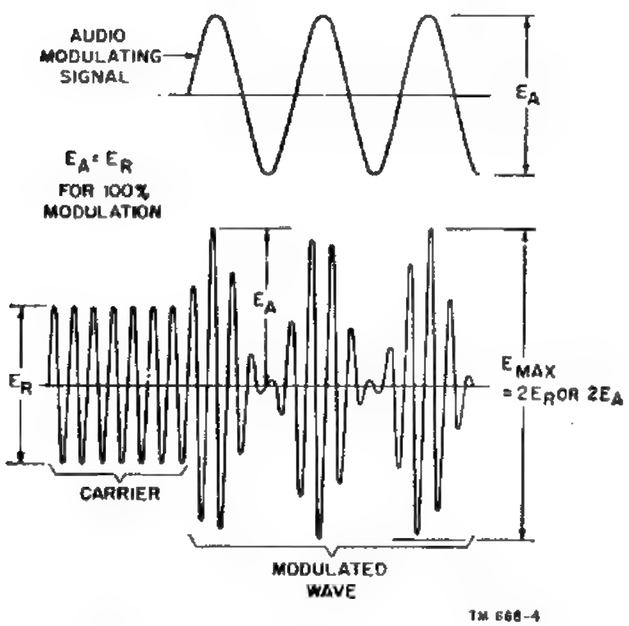


Figure 4. Carrier modulated 100 percent by a modulating signal.

(2) If E_A is less than E_R, percentages of modulation below 100 percent occur. If E_A is one-half E_R, the carrier is modulated only 50 percent (fig. 5). When the modulating signal swings to its maximum value in the positive direction, the carrier amplitude is increased by 50 percent. When the modulating signal reaches its maximum negative peak value, the carrier amplitude is decreased by 50 percent.

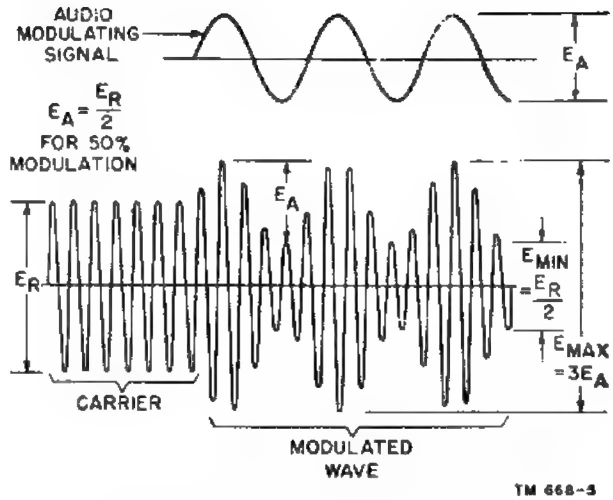


Figure 5. Fifty-percent modulation.

(3) It is possible to increase the percentage of modulation to a value greater than 100 percent by making E_A greater than E_R. In figure 6, the modulated carrier is varied from 0 to some peak-to-peak amplitude greater than $2E_R$. Since the peak-to-peak amplitude of the carrier cannot be less than 0, the carrier is cut off completely for all negative values of E_A greater than E_R. This results in a distorted signal, and the intelligence is received in a distorted form. Therefore, the percentage of modulation in a-m systems of communication is limited to values from 0 to 100 percent.

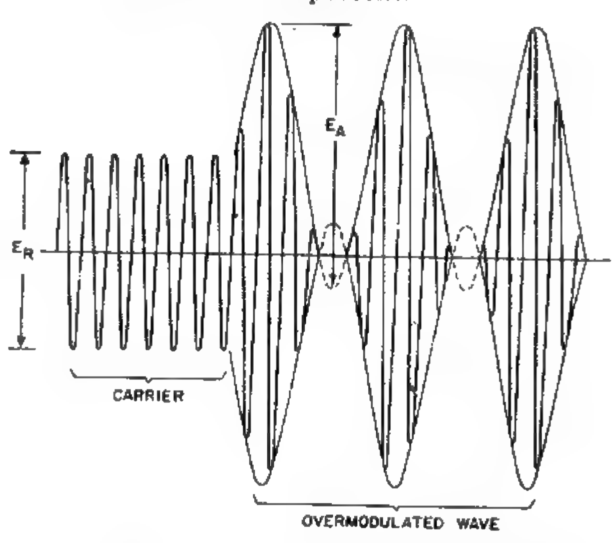


Figure 6. Overmodulation of carrier.

(4) The actual percentage of modulation of a carrier (M) can be calculated by using the following simple formula

$$M = \text{percentage of modulation} = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \times 100$$

where E_{max} is the greatest and E_{min} the smallest peak-to-peak amplitude of the modulated carrier. For example, assume that a modulated carrier varies in its peak-to-peak amplitude from 10 to 30 volts. Substituting in the formula, with E_{max} equal to 30 and E_{min} equal to 10,

$$M = \text{percentage of modulation} = \frac{30-10}{30+10} \times 100 = \frac{20}{40} \times 100 = 50 \text{ percent}$$

This formula is accurate only for percentages between 0 and 100 percent.

c. Side Bands.

(1) When the outputs of two oscillators beat together, or heterodyne, the two original frequencies plus their sum and difference are produced in the output. This heterodyning effect also takes place between the a-f signal and the r-f signal in the modulation process and the beat frequencies produced are known as *side bands*. Assume that an a-f signal whose frequency is 1,000 cps (cycles per second) is modulating an r-f carrier of 500 kc (kilocycles). The modulated carrier consists mainly of three frequency components: the original r-f signal at 500 kc, the sum of the a-f and r-f signals at 501 kc, and the difference between the a-f and r-f signals at 499 kc. The component at 501 kc is known as the *upper* side band, and the component at 499 kc is known as the *lower* side band. Since these side bands are always present in amplitude modulation, the a-m wave consists of a center frequency, an upper side-band frequency, and a lower side-band frequency. The amplitude of each of these is constant in value but the resultant wave varies in amplitude in accordance with the audio signal.

(2) The carrier plus the two side bands, with the amplitude of each component plotted against its frequency, is represented in figure 7 for the example given above. The modulating signal, f_A, beats against the carrier, f_C, to produce upper side band f_H and lower side band f_L. The modulated carrier occupies a section of the radio-frequency spectrum extending from f_L to f_H, or 2 kc. To receive this signal, a receiver must have r-f stages whose bandwidth is at least 2 kc. When the receiver is tuned to 500 kc, it also must be able to receive 499 kc and 501 kc with relatively little loss in response.

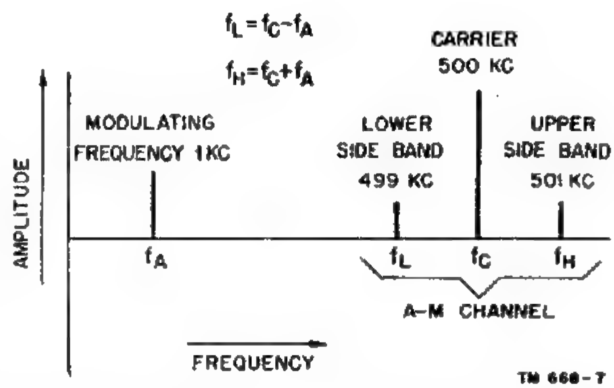


Figure 7. Frequency spectrum of amplitude-modulated wave.

(3) The audio-frequency range extends approximately from 16 to 16,000 cps. To accommodate the highest audio frequency, the a-m frequency channel should extend from 16 kc below to 16 kc above the carrier frequency, with the receiver having a corresponding bandwidth. Therefore, if the carrier frequency is 500 kc, the a-m channel should extend from 484 to 516 kc. This bandwidth represents an ideal condition; in practice, however, the entire a-m bandwidth for audio reproduction rarely exceeds 16 kc. For any specific set of audio-modulating frequencies, the a-m channel or band-

width is twice the highest audio frequency present.

(4) The r-f energy radiated from the transmitter antenna in the form of a modulated carrier is divided among the carrier and its two side bands. With a carrier component of 1,000 watts, an audio signal of 500 watts is necessary for 100-percent modulation. Therefore, the modulated carrier should not exceed a total power of 1,500 watts. The 500 watts of audio power is divided equally between the side bands, and no audio power is associated with the carrier.

(5) Since none of the audio power is associated with the carrier component, it contains none of the intelligence. From the standpoint of communication efficiency, the 1,000 watts of carrier-component power is wasted. Furthermore, one side band alone is sufficient to transmit intelligence. It is possible to eliminate the carrier and one side band, but the complexity of the equipment needed cancels the gain in efficiency.

d. Disadvantages of Amplitude Modulation. It was noted previously that random noise and electrical interference can amplitude-modulate the carrier to the extent that communication cannot be carried on. From the military standpoint, however, susceptibility to noise is not the only disadvantage of amplitude modulation. An a-m signal is also susceptible to enemy jamming and to interference from the signals of transmitters operating on the same or adjacent frequencies. Where interference from another station is present, the signal from the desired station must be many times stronger than the interfering signal. For various reasons, the choice of a different type of modulation seems desirable.

4. Phase Modulation

a. General.

(1) Besides its amplitude, the frequency or phase of the carrier can be varied to produce a signal bearing intelligence. The process of varying the frequency in accordance with the intelligence is frequency modulation, and the process of varying the phase is phase modulation. When frequency modulation is used, the phase of the carrier wave is indirectly affected. Similarly, when phase modulation is used, the carrier frequency is affected. Familiarity with both frequency and phase modulation is necessary for an understanding of either.

(2) In the discussion of carrier characteristics, carrier frequency was defined as the number of cycles occurring in each second. Two such cycles of a carrier are represented by curve A in figure 8. The starting point for measuring time is chosen arbitrarily, and at 0 time, curve A has some negative value. If another curve B, of the *same frequency* is drawn having 0 amplitude at 0 time, it can be used as a reference in describing curve A.

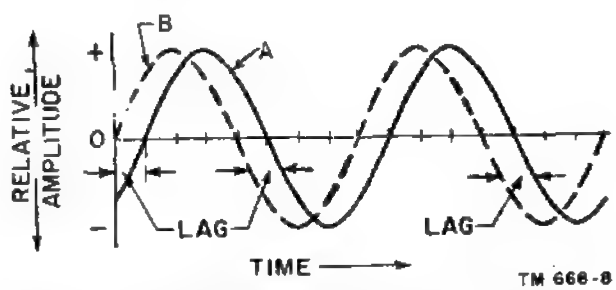


Figure 8. Determining relative phase from a curve of the same frequency.

(3) Curve B starts at 0 and swings in the positive direction. Curve A starts at some negative value and also swings in the positive direction, not reaching 0 until a fraction of a cycle after curve B has passed through 0. This fraction of a cycle is the amount by which A is said to *lag* B. Because the two curves have the same frequency, A will always lag B by the same amount. If the positions of the two curves are reversed, then A is said to *lead* B. The amount by which A leads or lags the reference is called its phase. Since the

reference given is arbitrary, the phase is relative.

b. Vector Representation.

(1) The cyclic changes of the carrier have been represented by plotting a curve of amplitude against time. It also is possible to represent the carrier cycle as the projection of a point rotating counterclockwise in a circle. This is called the vector representation and is accomplished by plotting the amplitude against the number of degrees of rotation of the point instead of directly against time. For each cycle of the carrier, the point rotates in one complete circle, or 360°. This is the *period* of the wave, or the time for one cycle.

(2) The carrier cycle as the projection of a point moving in a circle is plotted in figure 9. Starting from 0 the point rotates through 360° and back to 0, where the next cycle begins. When this point is projected along the set of axes on the right, one complete revolution of the point traces one complete cycle of the wave. The frequency of the wave in cycles per second is numerically equal to the revolutions per second made by the point. The relative phase of the wave shown is 0, since it has 0 amplitude at 0°. The peak amplitude of the wave is equal to the radius of the circle, and the peak-to-peak amplitude is equal to the diameter of the circle.

(3) The position of the point at any instant can be indicated by an arrow drawn from the center of the circle to the point. This arrow is called a *vector* and in the diagram the point is at a position of 45°. It must be remembered that this vector is not standing still, but is rotating at a frequency equal to that of the wave it represents. The vector is a convenient device with which to indicate phase relations between one wave and another or between a wave and an arbitrarily chosen reference. Suppose that a wave is observed staring at a time when its vector is in the 45° position (fig. 10). A second vector having 0 amplitude at 0 time and rotating in the same counterclockwise direction can be used as a reference for determining relative phase. In the graphic projection on the right, the two waves have the same frequency; however, the wave being observed is said to have a relative phase lead of 45°. An inspection of the figure shows that this is the central angle between the two vectors, as measured counterclockwise from the reference vector. The solid sine wave passes through 0 in a positive direction 45° ahead of the dotted reference sine wave shown at the right. If a particular carrier wave is considered in this manner, it has a relative phase relation at any instant to a reference carrier having the same frequency. The phase angle can be measured either in degrees or in radians. Since 360° equals 2π radians, the phase angle of 45° observed above can be expressed as $\pi/4$ radians, or simply $\pi/4$.

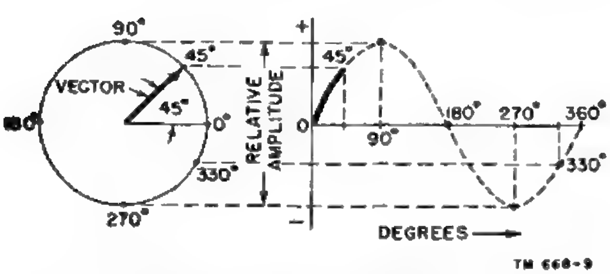


Figure 9. Projection of a point moving in a circle to form one cycle of a sine wave.

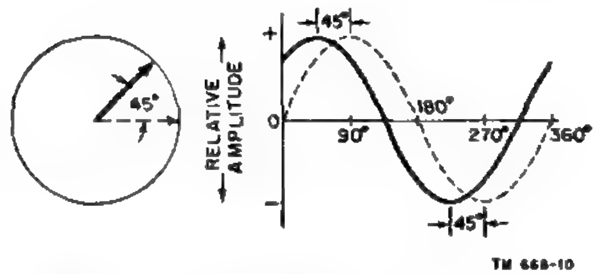


Figure 10. Measuring relative phase angle by means of vectors.

c. Phase Modulation.

(1) In phase modulation, the relative phase of the carrier is made to vary in ac-

cordance with the intelligence to be transmitted. The carrier phase angle, therefore, is no longer fixed. The amplitude and the average frequency of the carrier are held constant while the phase at any instant is being varied with the modulating signal (fig. 11). Instead of having the vector rotate at the carrier frequency, the axes of the graph can be rotated in the opposite direction at the same speed. In this way the vector (and the reference) can be examined while they are standing still. In A of figure 11 the vector for the unmodulated carrier is given, and the smaller curved arrows indicate the direction of rotation of the axes at the carrier frequency. The phase angle, ϕ, is constant in respect to the arbitrarily chosen reference. Effects of the *modulating signal* on the relative phase angle at four different points are illustrated in B, C, D, and E.

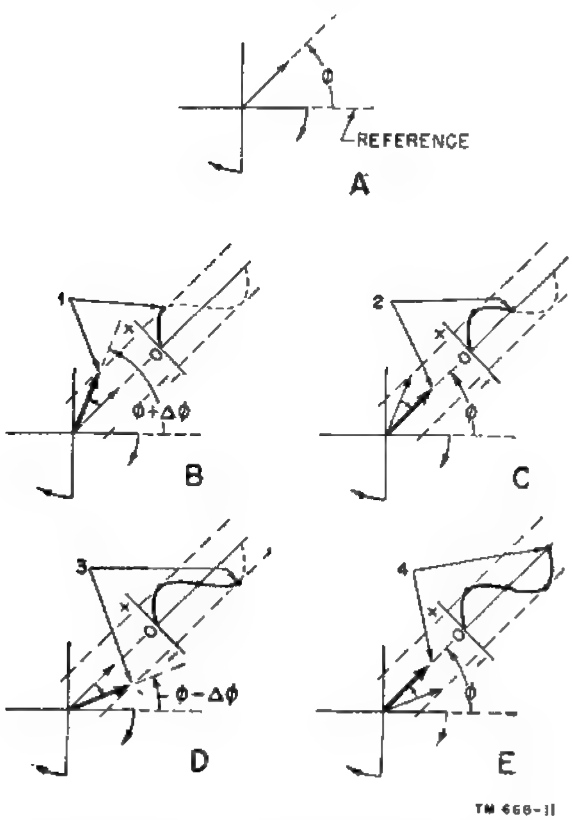


Figure 11. Successive vector representation of a phase-modulated carrier.

(2) The effect of a positive swing of the *modulating signal* is to speed the rotation of the vector, moving it counterclockwise and increasing the phase angle, ϕ. At point 1, the modulating signal reaches its maximum positive value, and the phase has been changed by the amount $\Delta\phi$. The instantaneous phase condition at 1 is, therefore, $(\phi + \Delta\phi)$. Having reached its maximum value in the positive direction, the modulating signal swings in the opposite direction. The vector speed is reduced and it appears to move in the reverse direction, moving toward its original position. When the modulating signal reaches 0 (position 2 in C), the vector has returned to its original position. The phase angle again is ϕ in respect to the reference. The modulating signal continues to swing in the negative direction and the vector is carried past its original position in a clockwise direction. When it reaches its maximum negative position, at 3 in D, the vector phase angle has been changed by $-\Delta\phi$, and the instantaneous phase angle is reduced to $(\phi - \Delta\phi)$. Finally, the vector returns to its original phase angle as the modulating signal amplitude falls to 0, at position 4 in E. The phase angle again is ϕ, and the cycle is complete. The entire cycle of phase shift is repeated for each cycle of modulating signal; that is, the frequency of the modulating signal is reproduced as the repetition rate of the cycle of phase shift.

(3) For each cycle of the modulating signal, the relative phase of the carrier is varied between the values of $(\phi + \Delta\phi)$ and $(\phi - \Delta\phi)$. These two values of instantaneous phase, which occur at the maximum positive and maximum negative values of modulation, are known as the *phase-deviation limits*. The upper limit is $+\Delta\phi$; the lower limit is $-\Delta\phi$. The relations between the phase-deviation limits and the carrier vector are given in figure 12, with the limits of $\pm\Delta\phi$ indicated.

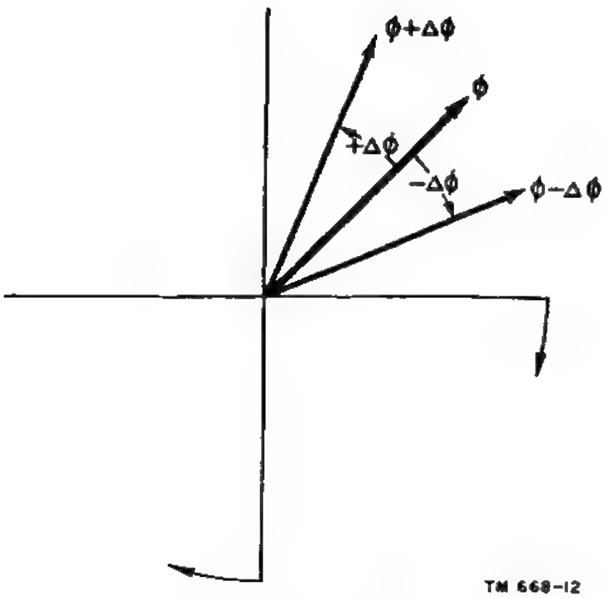


Figure 12. Phase-deviation limits of a modulated carrier.

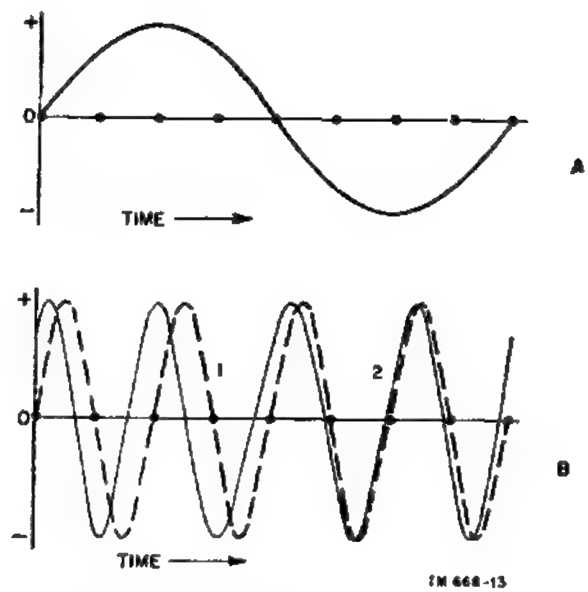


Figure 13. Phase-modulated carrier over 1 cycle of modulating signal.

(4) If the phase-modulated vector is plotted against time, the result is the wave illustrated in figure 13. The modulating signal is shown in A. The dashed-line waveform, in B, is the curve of the reference vector and the solid-line waveform is the carrier. As the modulating signal swings in the positive direction, the relative phase angle is increased from an original phase lead of 45° to some maximum, as shown at 1 in B. When the signal swings in the negative direction, the phase lead of the carrier over the reference vector is decreased to minimum value, as shown at 2; it then returns to the original 45° phase lead when the modulating signal swings back to 0. This is the basic resultant wave for sinusoidal phase modulation, with the amplitude of the modulating signal controlling the relative phase characteristic of the carrier.

d. P-M and Carrier Frequency.

(1) In the vector representation of the p-m carrier, the carrier vector is *speeded up* or *slowed down* as the relative phase angle is increased or decreased by the modulating signal. Since vector speed is the equivalent of carrier frequency, the carrier frequency must *change* during phase modulation. A form of frequency modulation, known as *equivalent f-m,* therefore, takes place. Both the p-m and the equivalent f-m depend on the modulating signal, and an instantaneous equivalent frequency is associated with each instantaneous phase condition.

(2) The phase at any instant is determined by the amplitude of the modulating signal. The instantaneous equivalent frequency is determined by the *rate of change* in the amplitude of the modulating signal. The rate of change in modulating-signal amplitude depends on two factors—the modulation amplitude and the modulation frequency. If the amplitude is increased, the phase deviation is increased. The carrier vector must move through a greater angle in the same period of time, increasing its speed, and thereby increasing the carrier frequency shift. If the modulation frequency is increased, the carrier must move within the phase-deviation limits at a faster rate, increasing its speed and thereby increas-

ing the carrier frequency shift. When the modulating-signal amplitude or frequency is decreased, the carrier frequency shift is decreased also. The faster the amplitude is changing, the greater the resultant shift in carrier frequency; the slower the change in amplitude, the smaller the frequency shift.

(3) The rate of change at any instant can be determined by the *slope*, or steepness, of the modulation waveform. As shown by curve A in figure 14, the greatest rates of change do not occur at points of maximum amplitude; in fact, when the amplitude is 0 the rate of change is maximum, and when the amplitude is maximum the rate of change is 0. When the waveform passes through 0 in the positive direction, the rate of change has its maximum positive value; when the waveform passes through 0 in the negative direction, the rate of change is a maximum negative value.

(4) Curve B is a graph of the rate of change of curve A. This waveform is leading A by 90°. This means that the

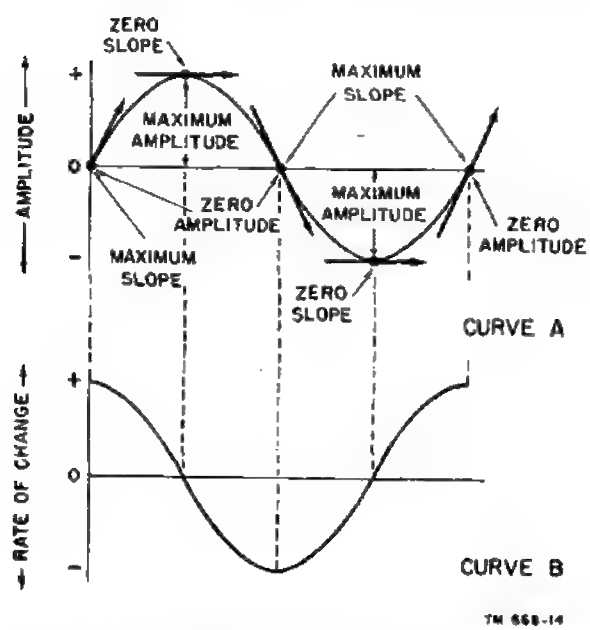


Figure 14. Amplitude compared to rate of change of a signal.

frequency deviation resulting from phase modulation is 90° out of phase with the phase deviation. The relation between phase deviation and frequency shift is shown by the vectors in figure 15. At times of maximum phase deviation, the frequency shift is 0; at times of 0 phase deviation, the frequency shift is maximum. The equivalent-frequency deviation limits of the phase-modulated carrier can be calculated by means of the formula,

$$\triangle F = \triangle\phi\, f \cos\,(2\pi \text{ ft})$$

where

$\triangle F$ is the frequency deviation,
$\triangle\phi$ is the maximum phase deviation,
f is the modulating-signal frequency,
cos (2π ft) is the amplitude variation of the modulating signal at any time, *t*.

When (2π ft) is 0 or 180°, the signal amplitude is 0 and the cosine has maximum values of +1 at 360° and —1 at 180°. If the phase deviation limit is 30°, or $\pi/6$ radians, and a 1,000-cps signal modulates the carrier, then

$$F = \frac{\pi}{6} \times 1{,}000 \times +1,$$

$$F = +523 \text{ cps, approximately.}$$

When the modulating signal is passing through 0 in the positive direction, the carrier frequency is raised by 523 cps. When the modulating signal is passing through 0 in the negative direction, the carrier frequency is lowered by 523 cps.

5. Frequency Modulation

a. When a carrier is frequency-modulated by a modulating signal, the carrier amplitude is held constant and the carrier frequency varies directly as the *amplitude* of the modulating signal. There are limits of frequency deviation similar to the phase-deviation limits in phase modulation. There is also an *equivalent phase* shift of the carrier, similar to the equivalent frequency shift in p-m.

b. A frequency-modulated wave resulting from 2 cycles of modulating signal imposed on

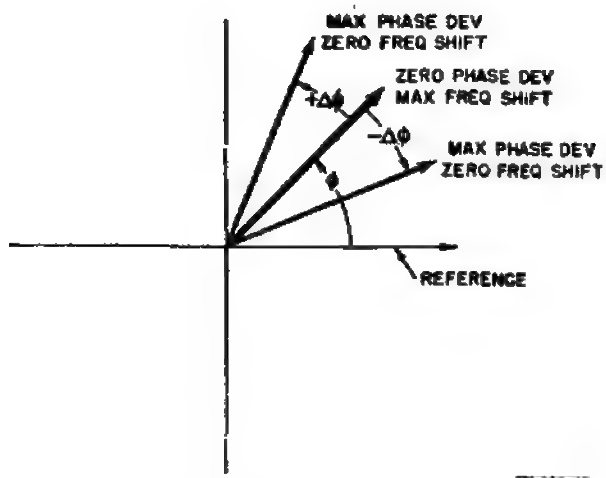


Figure 15. Phase deviation and frequency shift.

a carrier is shown in A of figure 16. When the modulating-signal amplitude is 0, the carrier frequency does not change. As the signal swings positive, the carrier frequency is increased, reaching its highest frequency at the positive peak of the modulating signal. When the signal swings in the negative direction, the carrier frequency is lowered, reaching a minimum when the signal passes through its peak negative value. The f-m wave can be compared with the p-m wave, in B, for the same 2 cycles of modulating signal. If the p-m wave is shifted 90°, the two waves look alike. Practically speaking, there is little difference, and an f-m receiver accepts both without distinguishing between them. Direct phase modulation has limited use, however, and most systems use some form of frequency modulation.

6. A-M, P-M, and F-M Transmitters

a. General. All f-m transmitters use either direct or indirect methods for producing f-m. The modulating signal in the direct method has a direct effect on the frequency of the carrier; in the indirect method, the modulating signal uses the frequency variations caused by phase-modulation. In either case, the output of the transmitter is a frequency-modulated wave, and the f-m receiver cannot distinguish between them.

b. A-M Transmitter.

(1) In the block diagram of the a-m transmitter (A of fig. 17), the r-f section consists of an oscillator feeding a buffer, which in turn feeds a system of frequency multipliers and/or intermediate power amplifiers. If frequency multiplication is unnecessary, the buffer feeds directly into the intermediate power amplifiers which, in turn, drive the final power amplifier. The input to the antenna is taken from the final power amplifier.

(2) The audio system consists of a microphone which feeds a speech amplifier. The output of this speech amplifier is fed to a modulator. For high-level modulation, the output of the modulator is connected to the final amplifier (solid arrow), where its amplitude modulates the r-f carrier. For low-

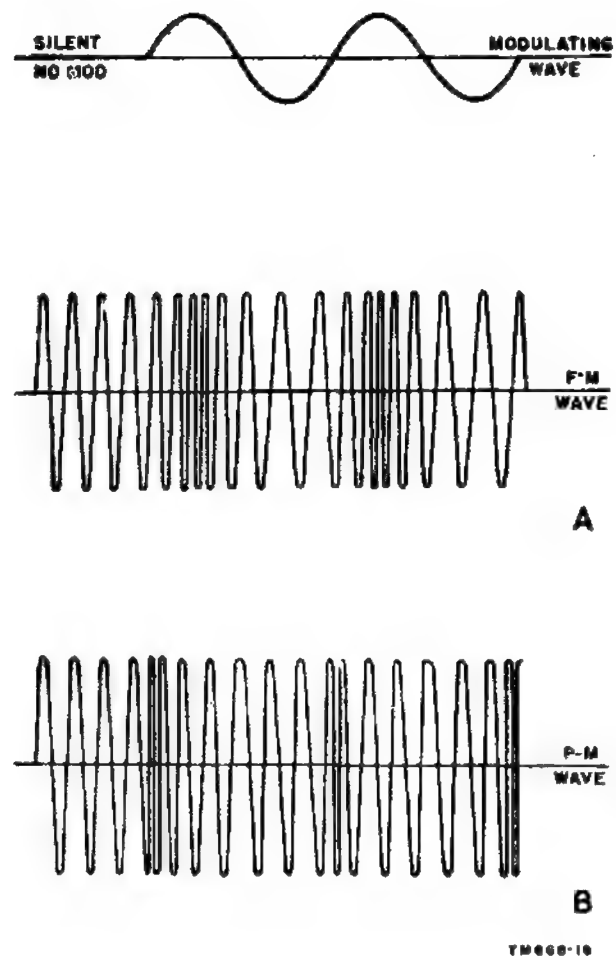


Figure 16. Comparison of f-m and p-m signals.

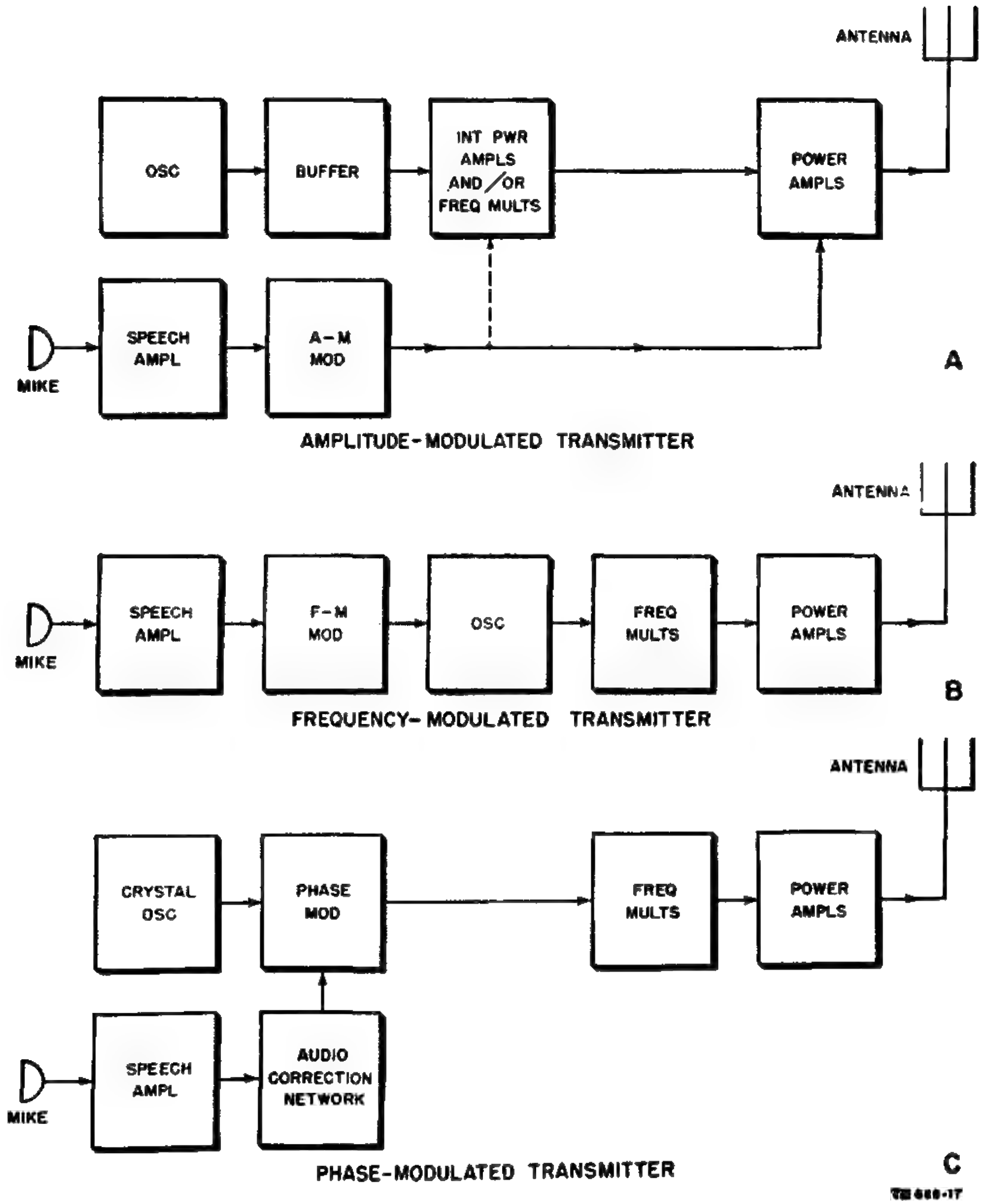


Figure 17. Basic a-m, p-m, and f-m transmitters.

level modulation, the output of the modulator is fed to the intermediate power amplifier (dashed arrow). The power required in a-m transmission for either high- or low-level modulation is much greater than that required for f-m or p-m.

c. P-M Transmitter. In the p-m, or indirect f-m, transmitter, the modulating signal is passed through some type of correction network before reaching the modulator, as in C. When comparing the p-m to the f-m wave, it was pointed out that a phase shift of 90° in the p-m wave made it impossible to distinguish it from the f-m wave (fig. 16). This phase shift is accomplished in the correction network. The output of the modulator which is also fed by a crystal oscillator is applied through frequency multipliers and a final power amplifier just as in the direct f-m transmitter. The final output is an f-m wave.

d. F-M Transmitter. In the f-m transmitter, the output of the speech amplifier usually is connected directly to the modulator stage, as in B. The modulator stage supplies an equivalent reactance to the oscillator stage that varies with the modulating signal. This causes the frequency of the oscillator to vary with the modulating signal. The frequency-modulated output of the oscillator then is fed to frequency multipliers which bring the frequency of the signal to the required value for transmission. A power amplifier builds up the signal before it is applied to the antenna.

e. Comparisons.

(1) The primary difference between the three transmitters lies in the method used to vary the carrier. In a-m transmission, the modulating signal controls the amplitude of the carrier. In f-m transmission, the modulating signal controls the frequency of the oscillator output. In p-m, or indirect f-m, transmission, the modulating signal controls the phase of a fixed-frequency oscillator. The r-f sections of these transmitters function in much the same manner, although they may differ appreciably in construction.

(2) The frequency multipliers used in a-m transmitters are used to increase the fundamental frequency of the oscillator. This enables the oscillator to operate at low frequencies, where it has increased stability. In f-m and p-m transmitters, the frequency multipliers not only increase the frequency of transmission, but also increase the frequency deviation caused by the modulating signal.

(3) In all three transmitters, the final power amplifier is used chiefly to increase the power of the modulated signal. In high-level a-m modulation, the final stage is modulated, but this is never done in either f-m or p-m.

7. A-M and F-M Receivers

a. General. The only difference between the a-m superheterodyne and the two basic types of f-m superheterodyne receivers (fig. 18) is in the detector circuit used to recover the modulation. In the a-m system, in A, the i-f signal is rectified and filtered, leaving only the original modulating signal. In the f-m system, the frequency variations of the signal must be transformed into amplitude variations before they can be used.

b. F-M Receiver. In the limiter-discriminator detector, in B, the f-m signal is amplitude-limited to remove any variations caused by noise or other disturbances. This signal is then passed through a discriminator which transforms the frequency variations to corresponding voltage amplitude variations. These voltage variations reproduce the original modulating signal. Two other types of f-m single-stage detectors in general use are the ratio detector and the oscillator detector, shown in C.

8. Summary

a. The carrier has the properties of frequency, amplitude, and relative phase.

b. In modulation, one of these three properties of the carrier is modified.

c. In amplitude modulation, the amplitude of the carrier is varied in proportion to the amplitude of the modulating wave.

d. The percentage of modulation of an a-m

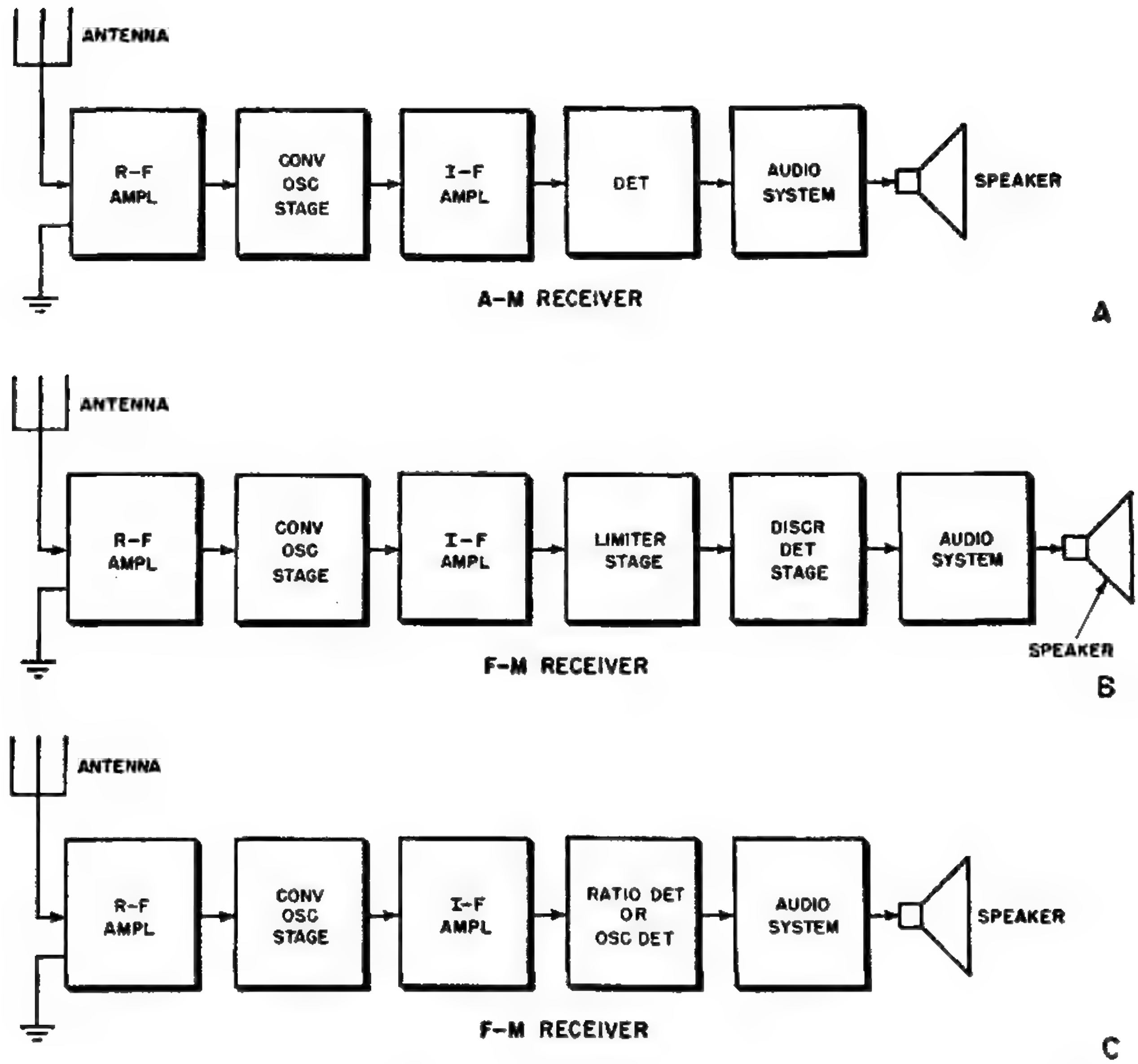


Figure 18. Basic a-m and f-m receivers.

wave is the percentage ratio of the peak-to-peak amplitude of the modulating voltage to the peak-to-peak amplitude of the carrier.

e. In amplitude modulation, the modulated wave consists of the carrier wave and of frequencies equal to the sum and difference between the carrier and the modulating frequency, called side bands.

f. The intelligence is contained in the side bands.

g. A-m has the disadvantage of being susceptible to some types of noise and interference.

h. In phase modulation, the instantaneous phase of the signal is varied by the modulating signal.

i. A change in phase is equivalent to an instantaneous change in frequency.

j. In a phase-modulation system, the equivalent frequency deviation is proportional to the frequency of the modulating signal.

k. When the carrier frequency is varied directly, the process is called direct f-m.

l. When the carrier frequency is varied indirectly, the process is called indirect f-m.

m. In a frequency-modulation system, the frequency varies directly with the amplitude of the modulating signal. The amplitude of the modulated wave remains constant, and the equivalent phase varies about a mean value.

9. Review Questions

a. What is meant by modulation of a carrier wave?

b. Name the most widely used types of modulation.

c. When a wave is amplitude-modulated 100 percent, what is the relationship between the amplitude of the modulating signal and that of the carrier?

d. What causes overmodulation of an a-m signal?

e. How many side bands are produced in a wave that is amplitude-modulated by a single sinusoidal tone?

f. What are the principal disadvantages of a-m?

g. Define phase deviation.

h. What happens to the instantaneous frequency during phase deviation?

i. How does the frequency of a p-m wave change during a cycle of modulating voltage?

j. How does the frequency of the f-m wave change during a single cycle of modulating voltage?

k. What characteristic of the modulating wave determines the maximum frequency deviation of an f-m wave?

CHAPTER 2

PRINCIPLES OF F-M

10. Frequency-Modulated Wave

a. Modulating Signal Amplitude and Frequency Deviation.

(1) In a frequency-modulated wave, the frequency varies instantaneously about the unmodulated carrier frequency in proportion to the amplitude of the modulating signal. When the modulating signal increases in amplitude, the instantaneous frequency increases, and when the modulating signal decreases, the frequency decreases.

(2) A radio-frequency carrier and an audio-frequency signal are shown separately in A and B of figure 19. When they are combined in the modulation process, the resultant signal is the f-m wave in C. As the amplitude of the audio signal increases in the positive direction, the modulated wave seems to bunch up, spreading out when the audio signal goes in the negative direction. These changes in the spacing of the modulated wave are caused by instantaneous changes in frequency. When the modulating signal is increased in amplitude, as in D, the changes in the spacing of the wave are proportionally greater, as in F. Therefore, the frequency deviation of the modulated wave is directly proportional to the amplitude of the modulating signal. When the audio voltage reaches its peak value in the positive direction, the frequency of the carrier is at its highest value above the center value. When the modulating voltage reaches its negative peak, the frequency of the carrier wave is reduced to its lowest value below that of the center carrier frequency. Maximum frequency deviation, therefore, takes place at the peaks of the audio signal.

b. Signal Frequency and Deviation. Figure 20 shows that each cycle of the modulating voltage, A, produces a corresponding variation in the frequency of the carrier wave, C. Two cycles of the audio wave produce 2 cycles of frequency change in the carrier. In D, the frequency of the modulating wave is increased so that in the *same* time interval the signal undergoes 3 complete cycles. Under this condition, the maximum frequency deviation remains the same as before; this is because the amplitude of the modulating wave has not been changed. The effective result of raising the frequency of the modulating signal is shown by comparing C and F. The audio signal is superimposed on the modulated carrier to demonstrate more clearly the relationship between the frequency of the modulating signal and the frequency changes in the carrier. This comparison indicates that, with a higher modulating frequency, the modulated wave deviates more frequently. However, the *limits* of frequency deviation are the *same,* regardless of the modulating frequency, since the audio signal is of constant amplitude.

c. Characteristics of F-M. The most important characteristics of a frequency-modulated wave are as follows: The amplitude of the modulated wave remains constant; the frequency of the modulated wave varies directly as the amplitude of the modulating signal; the limits of frequency shift on either side of the carrier are known as the *frequency deviation* limits. In an f-m system, the frequency of the modulating voltage determines the number of times per second that the frequency shifts between the

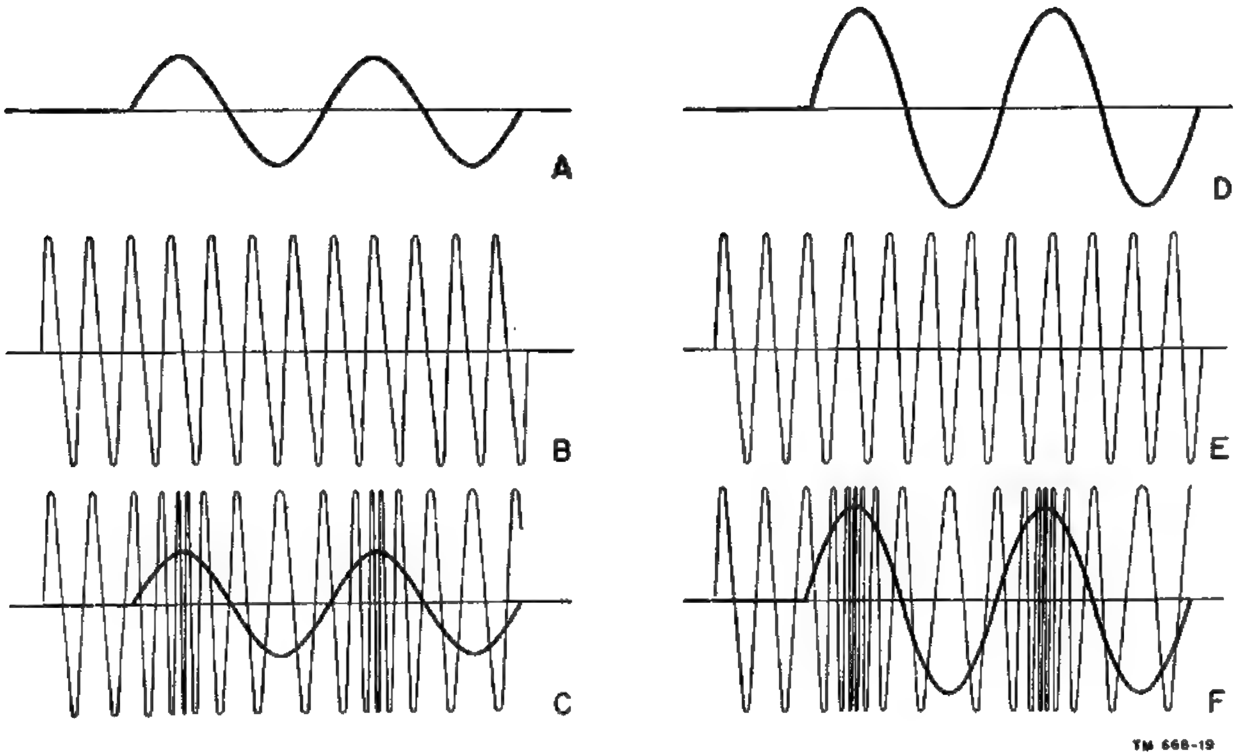


Figure 19. F-m waves for modulating signals differing in amplitude.

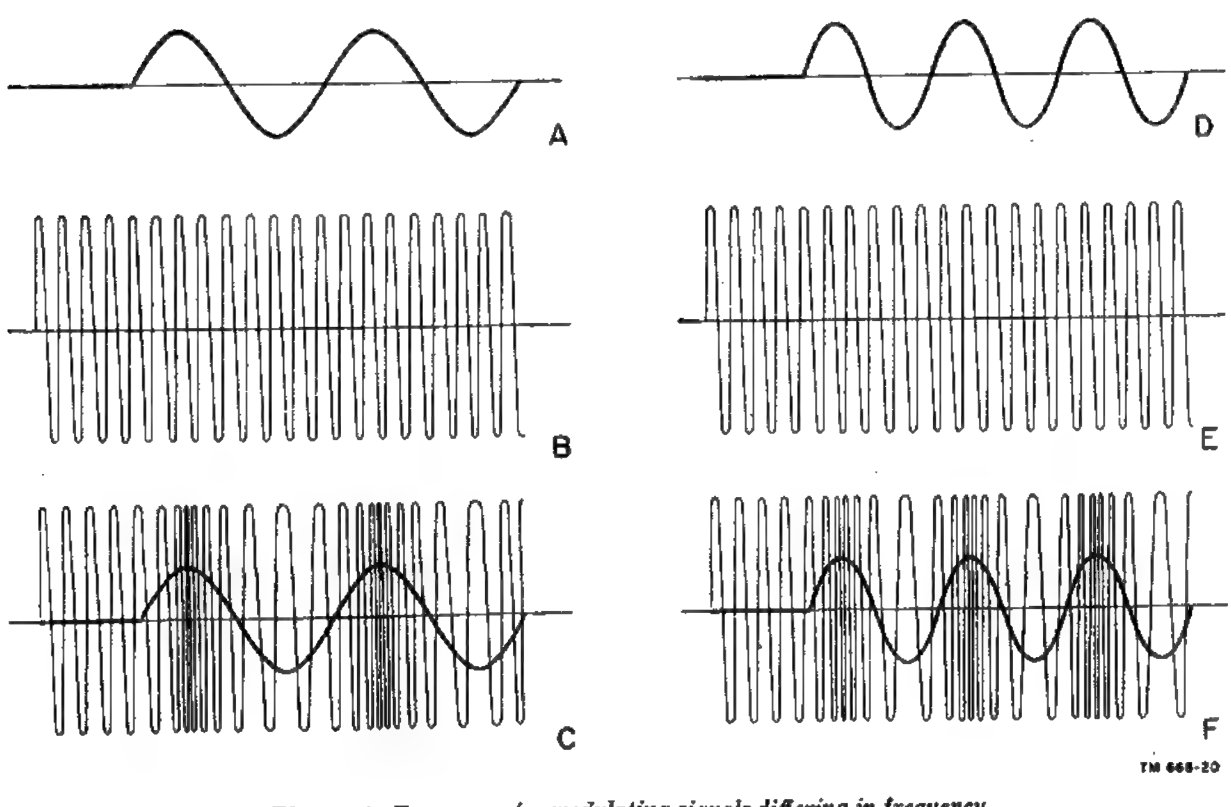


Figure 20. F-m waves for modulating signals differing in frequency.

deviation limits. The higher the frequency of the modulating signal, the greater number of times per second the frequency varies between the deviation limits set by the peak amplitude of the modulating signal. The ratio between the maximum frequency deviation and the maximum frequency of the modulating signal is called the *modulation index.*

$$\textit{Modulation index} = \frac{\textit{maximum frequency deviation}}{\textit{maximum frequency of modulating signal}}$$

d. Percentage of Frequency Modulation. The percentage of modulation of an f-m signal cannot be determined in the same manner as an a-m signal because 100-percent modulation would mean that the entire carrier varies in frequency from 0 to twice the carrier frequency. Percentage of modulation in f-m is defined as the percentage of maximum deviation incorporated in a transmitter for a particular type of service. For an f-m transmitter with maximum deviation of 75 kc, 100-percent modulation occurs when the transmitter deviates the full 75 kc. When the deviation falls to 37½ kc, the transmitter is being modulated only 50 percent. Such a definition is flexible, of course, and depends on the maximum deviation of the equipment used.

11. Side Bands

a. Amplitude Modulation.

(1) The intelligence superimposed on the carrier wave generates side-band frequencies closely adjacent to the carrier frequency. The generation of the side bands actually is the purpose of modulation, since the modulating energy cannot travel through space by itself. In an amplitude-modulated wave, the information represented by a sinusoidal modulating signal is carried in two side bands spaced on each side of the carrier frequency by an amount equal to the frequency of the modulating wave.

(2) If the modulating signal is composed of more than one sinusoidal wave, two side bands are generated for each sinusoidal component of the modulating wave. The greater the number of modulating frequencies present, the greater the number of side bands produced. However, only one pair of side bands is present for each frequency in the modulated wave, and the side bands farthest away from the carrier are those of the highest modulating frequency. Therefore, if the transmitter is amplitude-modulated no more than 100 percent, the bandwidth occupied by an a-m carrier and its side bands is twice the highest audio frequency in the modulating wave.

(3) In figure 21, frequency is plotted horizontally and the power contained in a signal of a particular frequency governs the vertical height. The carrier wave produced by the transmitter has only one frequency and its amplitude appears as a sharp line. Its power, represented by the heavy center line, is determined by the capabilities of the equipment. When a sine-wave signal from the audio power amplifier is combined with the carrier in the modulator stage, two identical side-band frequencies appear on either side of the carrier. The power of the applied audio signal is divided between them, and they are separated from the carrier by a frequency difference equal to the frequency of the audio signal. The carrier frequency of 1,000 kc is amplitude-modulated by a 5-kc tone, producing two side bands of equal power at 995 and 1,005 kc respectively, which are shown adjacent to the line representing the carrier. When a 10-kc modulating signal is added to the first, two additional side-band pairs appear, having frequencies of 990 and 1,010 kc, respectively.

b. Frequency Modulation.

(1) In an f-m wave, the amplitude of the modulating signal determines the departure of the instantaneous frequency from the center, or carrier frequency. The instantaneous frequency can be made to deviate as much as desired from the carrier frequency by changing the amplitude of the modu-

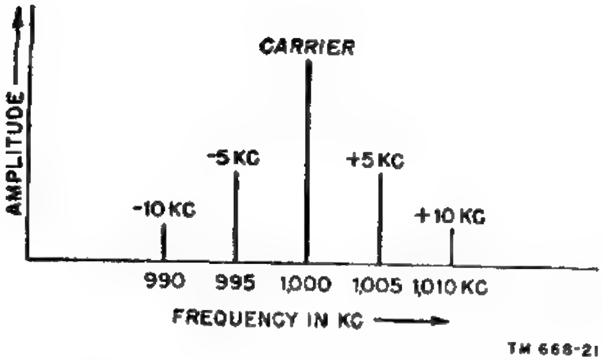


Figure 21. A-m carrier and side bands.

lating signal. It is possible to obtain a frequency deviation many times the frequency of the modulating signal itself. In practical equipment, this deviation frequency may be as high as several hundred kilocycles, even though the modulation frequency is but a few kilocycles. Therefore, the side bands generated by f-m are not restricted to the sum and difference between the highest modulating frequency and the carrier, as in a-m.

(2) Whereas in a-m, only two side bands, spaced equally on both sides of the carrier frequency, are generated, in f-m, many side bands are generated depending both in *number* and *amplitude* on the *modulation index*. The first pair of side bands in an f-m signal are those of the carrier frequency plus and minus the modulating frequency, and a pair of side bands will appear also at each multiple of the modulating frequency. As a result, an f-m signal occupies a greater bandwidth than does an a-m signal. For example, if a carrier of 1 mc (megacycle) is frequency-modulated by an audio signal of 10 kc, several side bands will be spaced equally on either side of the carrier frequency at 990 and 1,010, 980 and 1,020, 970 and 1,030, and so on. The total number present of *significant amplitude* (more than 1 percent of the amplitude of the unmodulated carrier) depends on the modulation index. With a high modulation index, more side bands are of appreciable amplitude, and the bandwidth is correspondingly greater.

c. Bandwidth. The maximum bandwidth of an amplitude-modulated transmission is twice that of the maximum frequency present in the modulating wave. Since the bandwidth in an f-m transmitter can exceed this by many times, the ratio of the bandwidth occupied to the absolute carrier frequency can be considerably larger than that of a-m transmitters. When a very wide bandwidth is used (wide-band f-m), it is necesary to choose a carrier frequency sufficiently high that the bandwidth is a small percentage of the carrier frequency, in order to permit a reasonable number of assigned channels. The f-m transmitter, however, can be adjusted so that the maximum bandwidth does not exceed that of an equivalent a-m transmission—that is, for the production of only one pair of side bands with significant amplitude. When the f-m transmitter is adjusted for a deviation that produces a bandwidth equal to that produced by an equivalently modulated a-m transmitter, it is called nfm (*narrow-band* f-m). With this narrow bandwidth, the transmitter can be operated at much lower frequencies (generally below 40 mc).

d. Relation Between Modulating Signal, Deviation, and Bandwidth. There are definite relations between the amplitude of the modulating signal, its frequency, the frequency deviation it produces, and the total bandwidth occupied by the resultant modulated f-m wave. If the frequency deviation is kept constant, *the number* of side bands *increases* as the modulating frequency *decreases*, and the total bandwidth occupied *decreases* as the modulating frequency *decreases*. The total bandwidth, however, can never be less than the bandwidth set by the peak-to-peak deviation alone, no matter how low the frequency of the modulating signal becomes. If the amplitude of the modulating signal *increases* and its frequency remains constant, the deviation *increases*, and the modulation index *increases*. This means that more energy goes into the outer side bands and correspondingly more of them increase to significant amplitude. The result is an increase in the number of useful side bands, as well as an increase in bandwidth.

e. Side-Band Spectrum and Modulation Index. The position of the side-band pairs for a single sinusoidal modulating wave depends only on the frequency of the modulating wave. The amplitude of the side bands depends on the ratio of the maximum frequency deviation of the carrier to the frequency of the modulating wave; that is, on the *modulation index.* The modulation index, in turn, depends on the amplitude of the modulating signal, because frequency deviation is proportional to signal amplitude. For a given modulation index and sinusoidal modulating frequency, the side-band pairs appear on either side of the carrier frequency (fig. 22). All the side-band components taken together form the *frequency spectrum* of the f-m wave. For example, if the modulating frequency is 15 kc and the frequency deviation is 75 kc, the modulation index will be 75/15, or 5, and the frequency components beyond the eighth pair of side bands will be less than 1 percent of the unmodulated carrier amplitude, and considered negligible.

f. Carrier Amplitude.

(1) The f-m wave consists of a center or carrier frequency and a number of side-band pairs, which, for a given audio frequency and amplitude, are constant. However, the resultant wave *varies in frequency* but is *constant in amplitude.* This resultant wave is the algebraic sum of the components which form it, and the carrier or center frequency will vary in amplitude with the modulation. When the transmitted signal is unmodulated, there is a certain constant amount of power in the carrier signal. When modulation is applied, power is taken from the carrier and forced into the side bands; therefore, the carrier amplitude, or center-frequency component, is reduced. The maximum power (fig. 22) is carried in the fourth side band, which is 4 times 15, or 60 kc, away from the carrier frequency.

(2) The carrier frequency or center-frequency component changes in amplitude with modulation, whereas, in a-m, the power for the side bands is supplied by the modulator and is not drawn away from the carrier. Since no information is in the carrier, reducing its amplitude increases the efficiency of operation in terms of power consumed. For some value of modula-

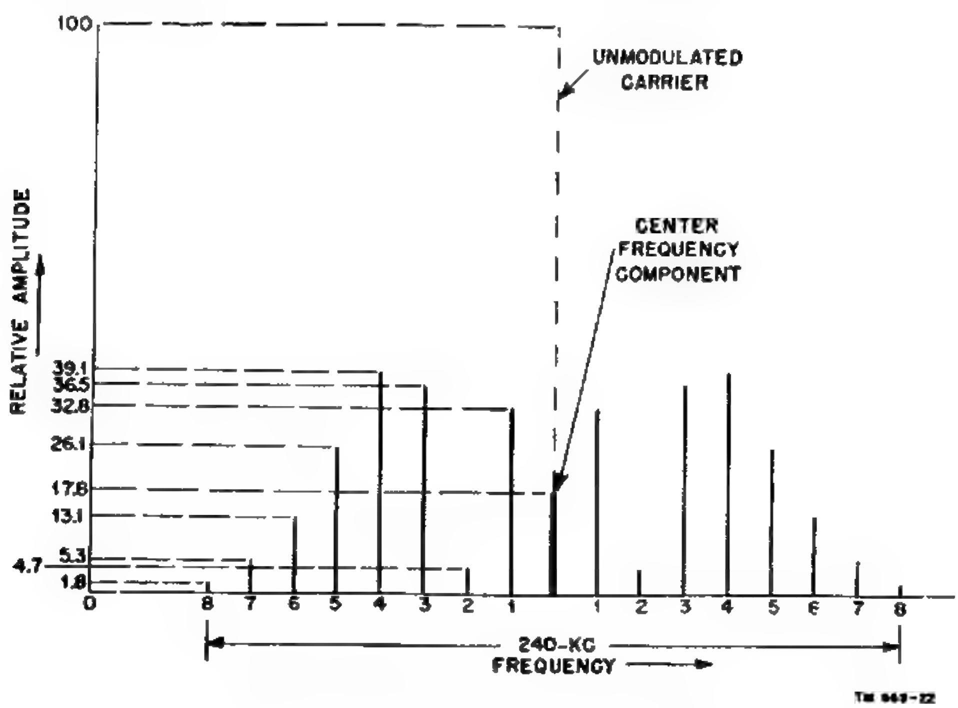


Figure 22. Frequency spectrum of f-m wave.

tion index and modulating frequency, carrier amplitude falls to 0 and all the power is contained in the side bands.

g. Numerical Values of Side Bands.

(1) In figure 22, with a frequency deviation of 75 kc and a modulating frequency of 15 kc, the center-frequency component is reduced to less than 20 percent of its unmodulated amplitude. If the modulating frequency is reduced to 5 kc with the same frequency deviation of 75 kc (fig. 23), the center frequency is reduced to only 1.4 percent of its unmodulated value. The side bands are spaced every 5 kc on either side of the center frequency out to the nineteenth pair of side bands, and all subsequent side bands are less than 1 percent of the unmodulated carrier amplitude.

(2) For the 15-kc modulating frequency of figure 22, with a modulation index of 5, the total bandwidth is 240 kc. In figure 23, with a 5-kc modulating frequency and a modulating index of 25, the total bandwidth occupied is 190 kc. The bandwidth of both is greater than the deviation limits of ±75 kc, which is equal to a peak-to-peak deviation of 150 kc. However, the side bands above or below the limit are relatively small in amplitude and can be disregarded. In both figures, the unmodulated carrier is shown in dashed lines for comparison with the amplitudes of the f-m side bands.

h. Side-Band Amplitude Computations.

(1) The relationship between the amplitudes of the side bands and an audio-modulating frequency for a modulation index of 2 is shown in figure 24. The deviation is 30 kc, with a modulating frequency of 15 kc. The modulated carrier wave, which is the resultant of the algebraic sum of the carrier and side bands is shown in A, with the modulating frequency, f_A, superimposed on it. At the positive peaks of the modulation cycle, the instantaneous frequency of the wave is f_C plus f_D, the frequency peak devia-

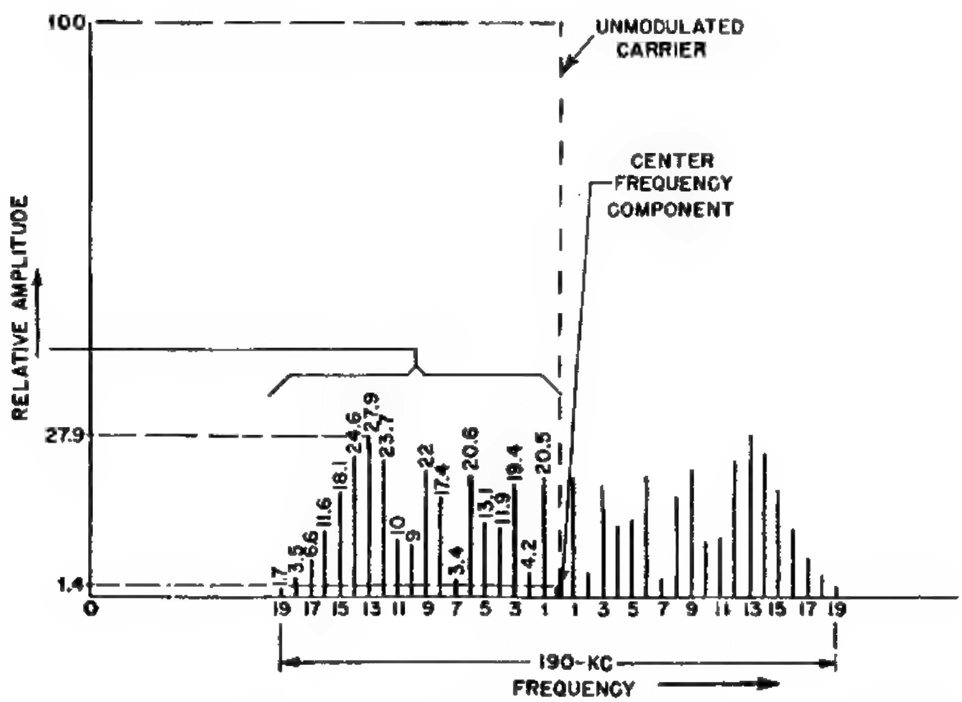


Figure 23. Another f-m frequency spectrum.

tion, whereas at the negative peaks the frequency is f_C minus f_D. The *peak-to-peak* deviation is, therefore, $2f_D$. If the carrier frequency is 100 mc with a frequency deviation of 30 kc, then the lower deviation limit is 99.97 mc and the upper one is 100.03 mc. There are four side-band pairs whose amplitudes exceed the 1-percent level and some of these are greater in amplitude than the center frequency. These are spaced on either side of the carrier frequency at intervals of 15 kc, as in C, D, E, and F, and each side-band pair has an amplitude as shown in G. The center frequency is reduced in amplitude to 22.4 percent of the unmodulated value. The first side-band pair at 99.985 mc and 100.015 mc has an amplitude of 57.7 percent of the unmodulated carrier value; the second side-band pair at 99.97 and 100.03 mc is 35.3 percent of the unmodulated carrier; and so on.

(2) To compute these bandwidths and side-band amplitudes, tables and graphs are available. Table I shows the number of *effective side-band pairs* for a given modulation index, and also the effective bandwidth that results for a given audio-modulating frequency, f_A. To compute the number of side-band pairs from the table, it is necessary to know the frequency deviation and the modulating frequency. The quotient of the two determines the modulation index, which in turn determines the number of effective side-bands pairs. For example, if the deviation is 25 kc and the modulating frequency is 5 kc, the modulation index is 5. From table I, a signal with a modulation index of 5 has eight effective side-band pairs, and the bandwidth is 16 times the modulating frequency of 5 kc, or 80 kc. The modulation indices that are less than .5 have only one pair of effective side bands and their effective bandwidths are all equal to $2f_A$. When using table I, if the modulation index is some fractional value, use the nearest

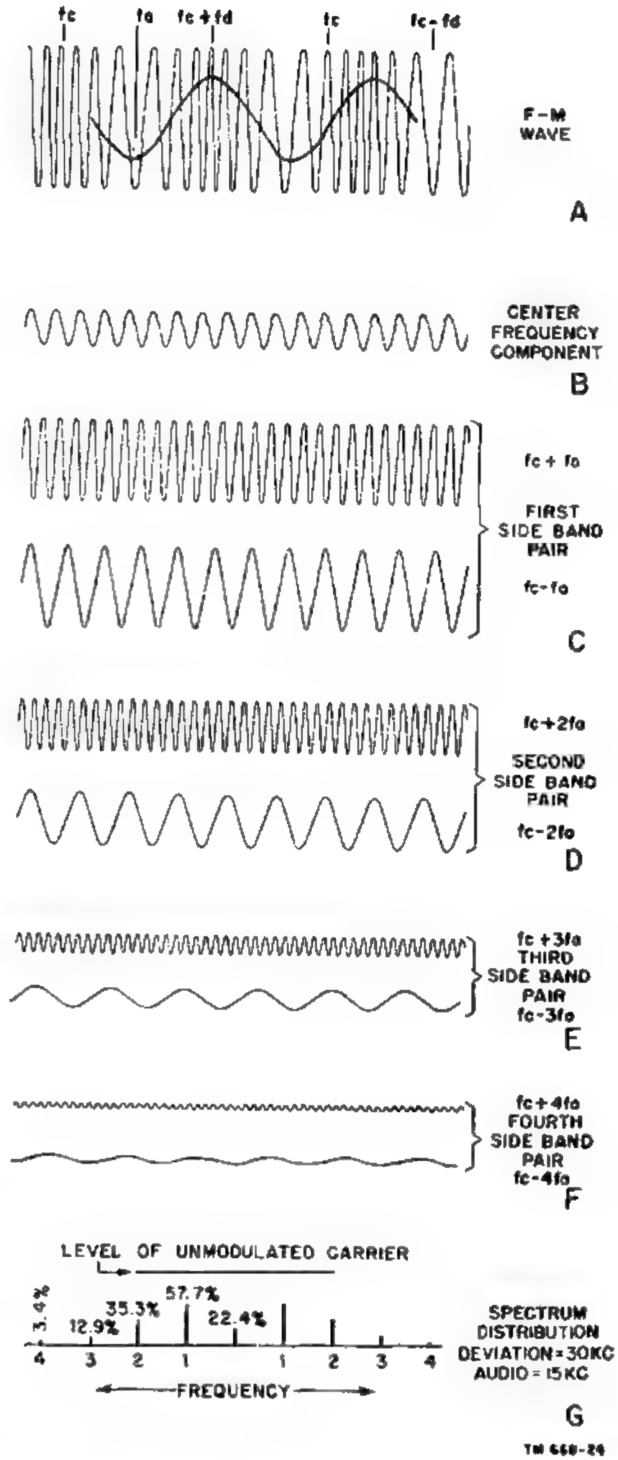


Figure 24. F-m carrier and side bands.

whole number. If the modulation index is $8\frac{1}{4}$, the number of effective side-band pairs for 8 is used; if it should be $8\frac{3}{4}$, the figure for 9 is used.

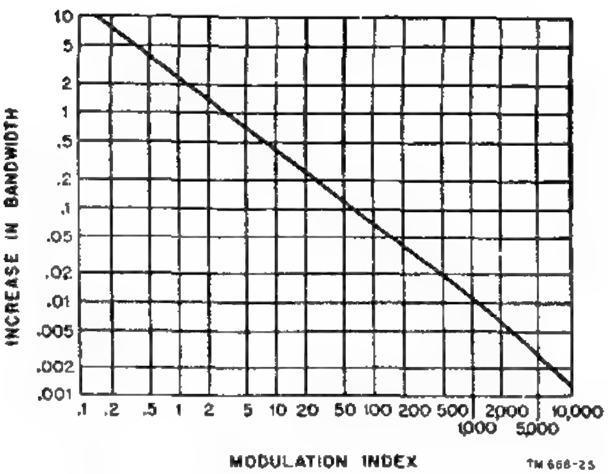


Figure 25. Bandwidth and modulation index.

Table I. Effective Bandwidth for Modulation Index

Modulation index $M = f_D/f_A$	Number of effective side-band pairs	Effective bandwidth
.5	2	$4f_A$
1	3	$6f_A$
2	4	$8f_A$
3	6	$12f_A$
4	7	$14f_A$
5	8	$16f_A$
6	9	$18f_A$
7	11	$22f_A$
8	12	$24f_A$
9	13	$26f_A$
10	14	$28f_A$
11	15	$30f_A$
12	16	$32f_A$
13	17	$34f_A$
14	18	$36f_A$
15	19	$38f_A$
16	20	$40f_A$
17	21	$42f_A$
18	23	$46f_A$
19	24	$48f_A$
20	25	$50f_A$
21	26	$52f_A$
22	27	$54f_A$
23	28	$56f_A$
24	29	$58f_A$
25	30	$60f_A$

(3) The same relationship is shown graphically in figure 25. The modulation index is plotted horizontally and the *increase* in bandwidth over the peak-to-peak deviation is shown vertically. A modulation index of 5 produces an increase of bandwidth of .6 or 60 percent. The peak-to-peak deviation is 2 times 25, or 50 kc. Sixty percent of 50 kc is 30 kc, and 30 kc plus 50 kc is equal to 80 kc. This is the same value as obtained with table I. The graph is especially useful in finding fractional values and modulation indices of less than one-half and more than 25.

i. Bandwidth and Guard Bands. The effective side bands must be at least as far from the carrier as the frequency deviation limits. Because of this, it is necessary to provide a channel or bandwidth that will handle the highest side-band component, plus a *guard band* that will absorb any side bands that extend beyond these limits. Since the modulating signal cannot always be specified and can vary over wide limits, it is easier to assign the channel in terms of deviation limits, and then to set aside some additional frequency space for guard bands on either side of the deviation limits. The channel with the guard bands assures a minimum amount of interference to stations operating on adjacent frequencies. In addition, the center carrier frequency of a particular station may vary from the assigned value and cause interference to stations on adjacent channels. The guard bands help to prevent this.

12. Types of Modulating Signals

a. Nonsinusoidal Modulating Waves. The preceding paragraphs are based on the assumption that the carrier wave was modulated by only one audio frequency that was sinusoidal in form. In general, waves containing information are neither sinusoidal nor composed of only one frequency. Moreover, the waves need not even be continuous; that is, they can start or stop abruptly. The modulation of the carrier by a nonsinusoidal wave can be understood by analyzing the character of the modulating wave. Waveforms can be analyzed as the sum of a number of sine waves of various amplitudes and frequencies. In respect to the frequency modulation of a carrier by a nonsinusoidal, discontinuous wave, the effective frequency variations follow the variations in the amplitude of the wave, just as they do in the sinusoidal wave. Therefore, the frequency deviation is still proportional to the peak amplitude of the signal,

although the variations are not as regular in a given time interval as they are in sine-wave modulation.

b. Speech Modulation. In normal speech, many different frequencies are present at the same time in the equivalent electrical wave. These frequencies range from 100 to 8,000 cps. For good intelligibility of speech only those between 300 and 3,000 cps need be transmitted. The average frequency content of speech depends on the voice of the speaker and on what he is saying; therefore, it is impossible to predict the side bands accurately. Standards must be set up that will duplicate the average speech spectrum, and a combination of 300 to 3,000 cps generally is used. The energy in high-pitched speech tones is much lower than in low-pitched speech, the amplitude of the 3,000-cps component being only one-tenth of the 300-cps components in a normal male voice. Therefore, the level of the 3,000-cps tone must be lower in amplitude to simulate the real qualities of speech. This means that the deviation for the high frequencies is correspondingly less, and the bandwidth is less than expected if it is assumed that equal intensities are used at 300 and 3,000 cps. Therefore, the bass frequencies cause the greatest deviation, even though the higher frequencies, if of equal intensity, tend to cause a wider bandwidth.

c. Pulse Modulation.

(1) When amplitude modulation with two audio frequencies is superimposed on the carrier, a pair of side bands is produced for every sine-wave component in the modulated wave. This fact can be demonstrated for as many components as are necessary to make up any given waveshape. A of figure 26 shows a rectangular wave composed of a large number of sine waves of various amplitudes and frequencies added together. Similarly, any other nonsinusoidal waveform can be analyzed in terms of a number of component sine waves.

(2) When nonsinusoidal waveforms amplitude-modulate a signal, they produce the same side bands that would be present if all of the sine waves to which they are equivalent were present. For example, rectangular amplitude modulation of a carrier wave produces side bands in symmetrical pairs about the carrier in relation to the amplitudes and frequencies of the sine-wave equivalent of the modulating waveform. This produces the spectrum of side bands shown in B. Each component making up the wave in A is plotted in terms of frequency and amplitude. This type of signal frequently is designated under a separate category as *pulse modulation.* Intelligence can be transmitted in such a system by controlling the amplitude,

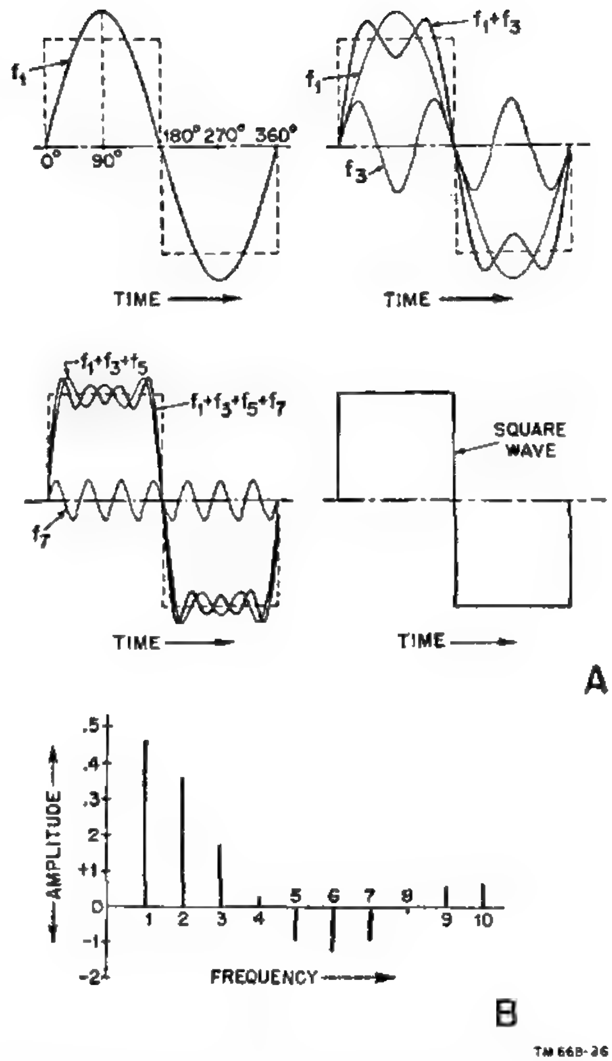


Figure 26. Sine-wave components of rectangular wave.

width, spacing, and position of the pulses. Frequency modulation, with the modulating wave in the form of rectangular pulses, is used in *frequency-shift* telegraphy. The frequency of the carrier is shifted abruptly between two values in accordance with the rectangular pulses received from a teletypewriter. Triangular pulses derived from certain types of facsimile transmitters for picture transmission are used also.

d. Side Bands—Complex Waves.

(1) When a sine-wave frequency modulates a carrier, many side bands are produced on either side of the carrier. These side bands are simple multiples of the modulating frequency with upper and lower side bands being identical in amplitude and *symmetrical* about the carrier.

(2) Assume that a signal of 5 kc modulates a carrier wave and produces 10-kc deviation, while at the same time a 10-kc signal modulates the carrier and produces a 10-kc deviation. The spectrum produced by the two signals tends to interact in a complicated way, depending on the amplitude and phase of the various side-band components. Not only the spectrum of each component is produced, but all possible combinations of each component with every other are present. Not all of these combinations, however, are of significant amplitude and the result is the production of the spectrum shown in figure 27. The actual instantaneous frequency deviation of the signal is the resultant of the sum of the two waves. That is, the two modulating waves add vectorially to produce a third that actually modulates the carrier, as shown in B.

(3) It can be seen that this spectrum is symmetrical about the carrier frequency but no longer corresponds to the spectrum of either component alone. Assume, however, that the deviation produced by the two component waves is not the same. In figure 28, a 5-kc signal produces a 30-kc deviation, and a 10-kc signal produces a 10-kc deviation. The modulating wave is shown with each of its components in B. The dotted line represents a 5-kc wave at three times the amplitude of the 10-kc wave. Since the deviation is proportional to the amplitude of the modulating signal, the amplitude of the wave producing 30-kc deviation is three times that of the wave producing 10-kc deviation. The result of combining the two waves is shown by the heavy black line, and it is this wave that causes the actual frequency deviation. The resultant spectrum is shown in A, and it is not symmetrical in respect to the carrier.

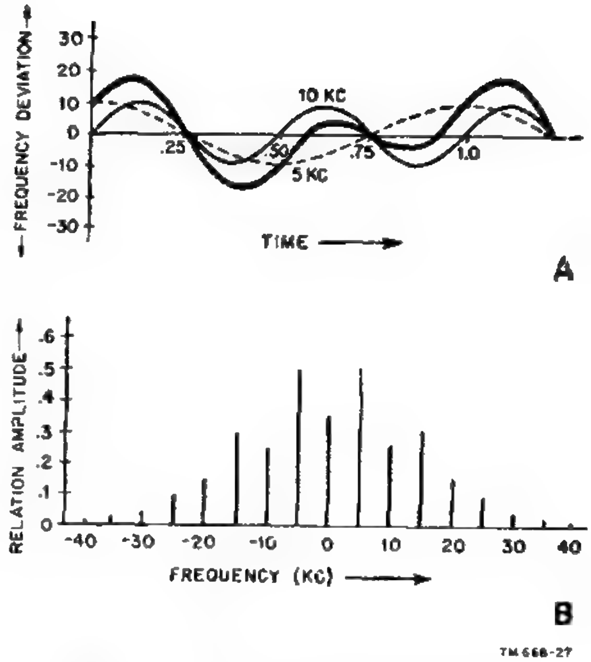


Figure 27. F-m for two modulating signals.

(4) In general, for a modulating signal that contains more than one sine wave, the side-band spectrum is not symmetrical about the carrier unless the modulating signal is symmetrical about the horizontal axis. For example, the wave (heavy line) in A of figure 27 is the same above and below the axis. This is not true of the wave in B of

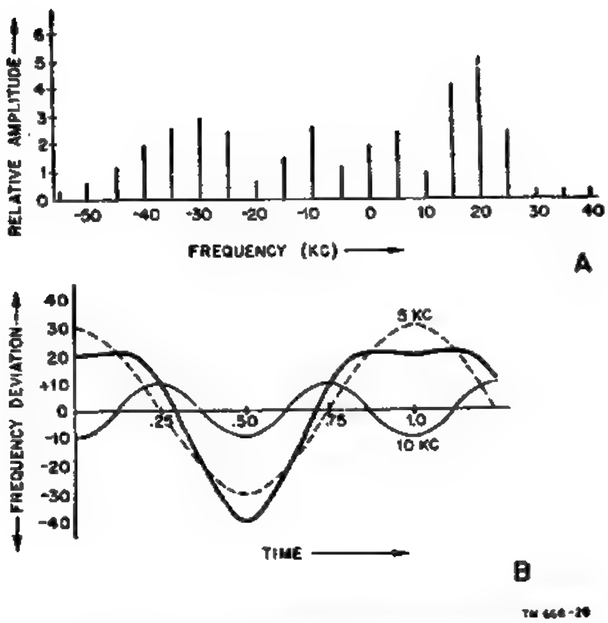


Figure 28. F-m for signals producing different deviations.

figure 28. The upper side band is associated with the shape of the positive half of the wave, whereas that of the lower side band depends on the negative half. This stems from the basic fact that an increase in amplitude of the modulating signal produced an increase in the instantaneous frequency of the carrier. Moreover, the energy in the side bands is distributed in accordance with the shape of the modulating wave. The effective bandwidth of the signal is increased, depending on whether the addition of the equivalent component sine waves of the complex waveform increases or decreases the peak deviation of the signal.

(5) Using the graph shown in C of figure 29, it is possible to determine the bandwidth of the signal with pulse modulation. For a wave with rectangular frequency modulation, shown in A, the spectrum is that in B. The maximum bandwidth required depends on the modulation index. This chart is used in the same manner as figure 25. When the deviation and the repetition rate of the rectangular pulses are known, the modulation index can be found by dividing the deviation by the repetition rate.

(6) For example, a frequency-shift telegraph transmitter deviation is 10 kc. (Each pulse of modulating signal shifts the frequency abruptly 10 kc). If the pulse repetition rate is 1,000 cycles per second, the modulation index is equal to

$$\frac{\textit{deviation}}{\textit{pulse repetition rate}} = \frac{10 \text{ kc}}{1 \text{ kc}} = 10$$

Entering the horizontal axis at a modulation index equal to 10, intersect with the curve at an increase in bandwidth of 2.25. The peak-to-peak deviation is 10 times 2 = 20 kc. The increase in bandwidth is, therefore, 20 times 2.25 = 45 kc, and the total band-

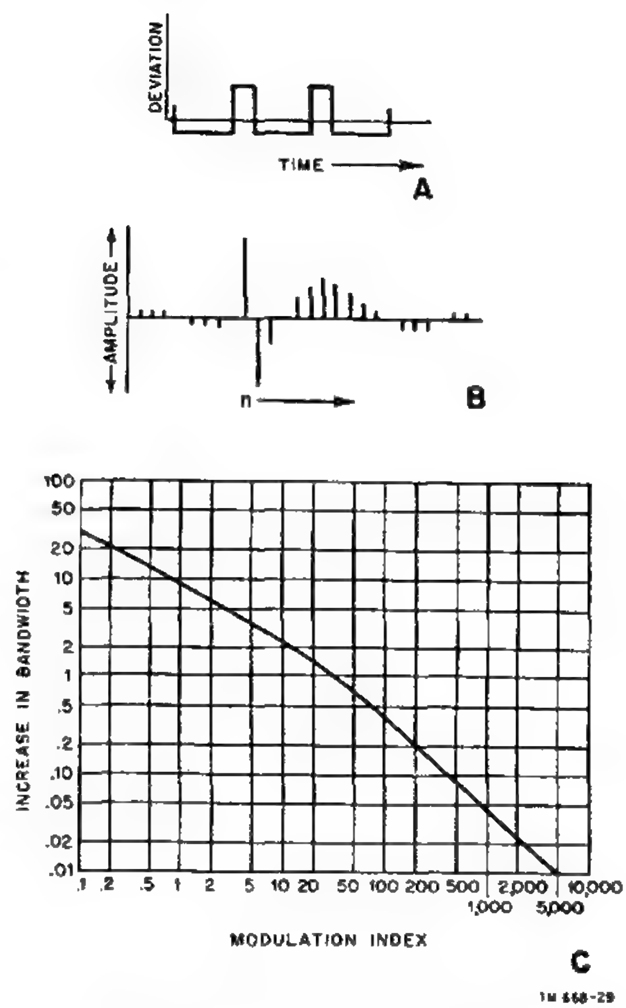


Figure 29. Rectangular modulation.

width is 20 plus 45 = 65 kc. The chart is computed with the pulse present one-quarter of the total time.

(7) If the modulating wave is triangular, as shown in A of figure 30, it produces the spectrum shown in B, and the total bandwidth can be calculated from the curves in C. This chart is used in the same manner as the one for the rectangular pulse. The modulation index is the quotient of the deviation and the repetition rate.

e. Bandwidth Limits for Nonsinusoidal Speech Modulation.

(1) The wide-band f-m wave contains numerous side bands that are distributed on either side of the carrier frequency. For sinusoidal modulating waves, the side bands are distributed equally above and below the carrier frequency. However, since most speech waves are not sinusoidal, the side bands in general are not of equal amplitude above and below the carrier. The number of side bands produced by a speech wave exceeds the number of sinusoidal components by a considerable amount. Moreover, the amplitude and the phase of each of these side bands vary in relation to the carrier frequency. Beyond certain frequency limits, however, the power contained in the outer side bands becomes very small, and so for all practical purposes they can be neglected. The point below which the side-band energy is negligible is arbitrary, depending on the kind of equipment that must be operated in the adjacent channel. As with sinusoidal modulation, when the the amplitude of the side-band component is less than 1 percent of the unmodulated carrier value, it is considered negligible.

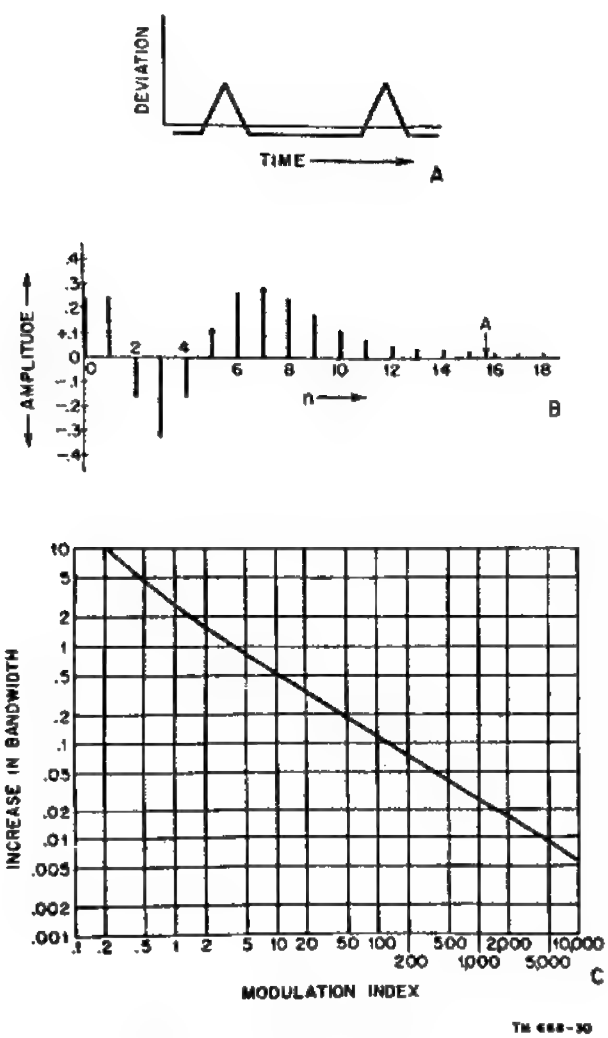


Figure 30. Triangular modulation.

(2) When the modulating signal is a mixture of many different waves distributed throughout the audio range, the ratio between frequency and phase deviation is hard to determine. The ratio of peak deviation in frequency to peak phase swing (*swing ratio*), measured in radians, gives some idea of the different conditions that prevail. This ratio, which is a constant for a simple sinusoid with a given modulation index, varies considerably with speech waves, depending on whether the modulation is indirect f-m or direct f-m. It is also dependent on the microphone, the characteristics of the audio-frequency amplifier, and how the total modulation applied to the transmitter is controlled to prevent overmodulation. In general, the swing ratio is greater than unity in an indirect f-m transmitter, whereas in a

direct f-m transmitter it is less than or equal to one.

(3) This means that in direct f-m transmitters the swing, or deviation in kilocycles, can be *less* for a given phase deviation than the expected value for a sine wave when speech is used. Conversely, the frequency deviation in an indirect f-m transmitter is *more* than the sine-wave value by as much as 50 percent, all depending on the character of the microphone, the audio amplifier, modulation limiting devices, and so forth. The importance of this relationship is that a directly modulated f-m transmitter has less peak deviation on speech waves than an indirect f-m transmitter when set for the same deviation, using a sine-wave source.

13. Preemphasis and Deemphasis

a. Preemphasis.

(1) In the transmitters used to convey speech, the deviation is the same for a given amplitude regardless of the frequency of the modulating signal. However, as signals pass through the transmitter, the receiver, and the space between them, certain amounts of unwanted noise and distortion are superimposed on the desired speech. This noise is distributed uniformly throughout the audible spectrum. Therefore, the ratio of the signal to the unwanted noise decreases in the higher frequencies because the speech amplitudes in this range do not have the intensity that the lower frequencies have. Moreover, the distortion increases in the high-frequency portion of the spectrum. The high frequencies make the greatest contribution to intelligibility of speech waves, since the consonants, which form the majority of speech sounds, have their peak energy in this part of the audio band.

(2) To avoid degrading the reproduction of consonants through poor signal-to-noise ratio in the upper end of the spectrum, a certain amount of added amplification (preemphasis) is provided for these frequencies. The result of this process should not sound unnatural when received, and the reverse procedure, *deemphasis,* therefore is used at the receiver. This combination of preemphasis and deemphasis provides a more uniform signal-to-noise ratio throughout the audio range. A transmitter using preemphasis has a wider side-band spectrum for speech than one without it. In general, the bandwidth of a speech signal deviating a transmitter 100 percent with preemphasis is about one-third greater than the deviation limits. If the deviation is 75 kc, for example, the total bandwidth at 100-percent modulation is about 200 kc (150 plus 50).

(3) The fact that preemphasis results in a greater bandwidth for a given deviation always must be taken into account. However, the possibility of overmodulation is not likely, since the high-frequency components of the signal originally are weak and the preemphasis merely brings them up to the level of the low tones. It does not cause overmodulation of an f-m transmitter, although the deviation limits set for the particular unit will be increased. It has been shown that the effective bandwidth increases as the audio frequency increases, and also that, as the upper audio frequency increases in level, more and more of the outer side bands rise above the 1-percent margin.

(4) The preemphasis characteristic of an f-m transmitter can be specified by a graph (fig. 31) showing the relationship between the audio input and the modulated output. The frequency of the audio spectrum is plotted horizontally, and the output of the unit for an input that is constant in respect to frequency is shown vertically. This curve shows that the output remains relatively constant from 50 to about 500 cps and then rises abruptly to a peak at 15,000 cps. Since this rise is specified in db (decibels), a change of

6 db means a doubling of the amplitude of the signal. Therefore, when the graph shows a rise of 18 db from 1,000 to 15,000 cps, it means that the amplitude has doubled three times. The resultant output at 15 kc, therefore, is 2 times 2 times 2, or 8 times the output at 1,000 cps.

b. Deemphasis. At the receiver, the reverse characteristic of preemphasis is used so that the natural balance between high and low frequencies in speech is not upset. The characteristics of preemphasis and deemphasis normally are achieved by simple electrical combinations of resistance, capacitance, and inductance connected to give the desired relationship between the input and the output voltages of the network. The characteristics of speech are complicated, and, therefore, the networks chosen represent a compromise between duplicating the exact loss of high frequencies and using as few parts as possible. In general, preemphasis and deemphasis circuits are very simple combinations of a capacitor and a resistor, or an inductor and a resistor.

c. Preemphasis Network. A simple preemphasis network consisting of an inductor and a resistor connected in the grid circuit of a vacuum-tube amplifier is shown in A of figure 32. In this circuit the audio voltage is impressed across the inductor and the resistor in series, and the output is taken across the inductor. Since the impedance of the coil rises with frequency, and the resistance remains constant, the voltage across the coil rises. The ratio of inductance to resistance determines the time constant of the combination, and the preemphasis characteristic can be specified completely in terms of the time constant. When the inductance is given in henrys and the resistance in megohms, the time constant is in microseconds. For example, calculate the time constant of the network (fig. 32) which contains a resis-

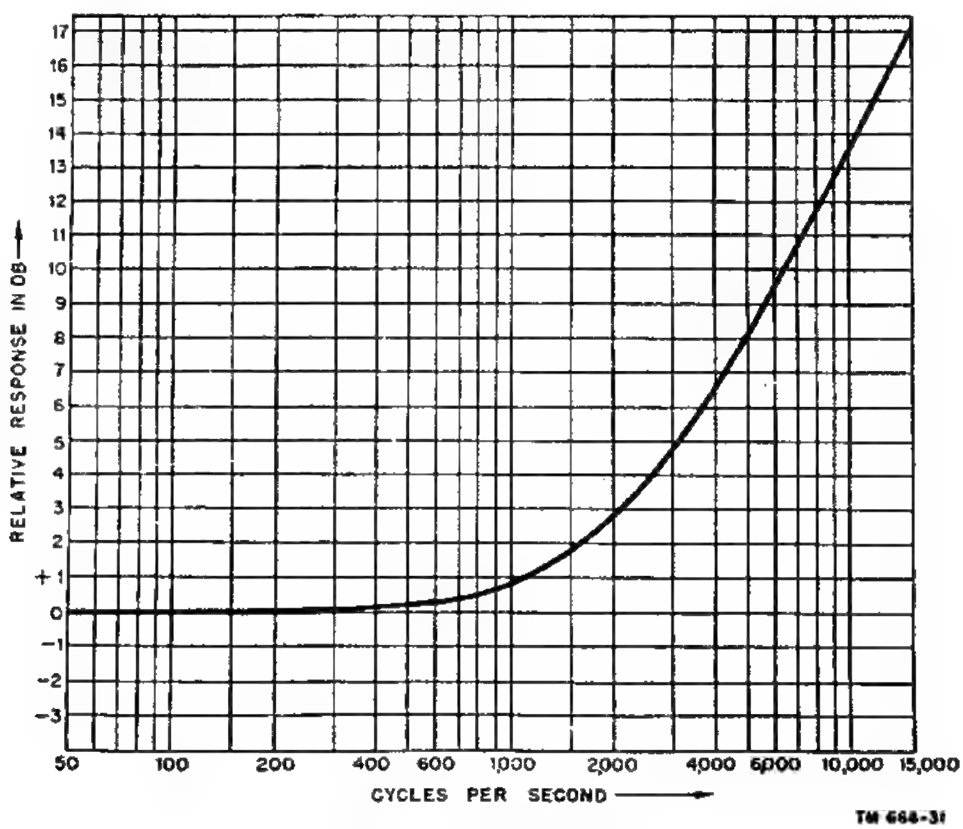
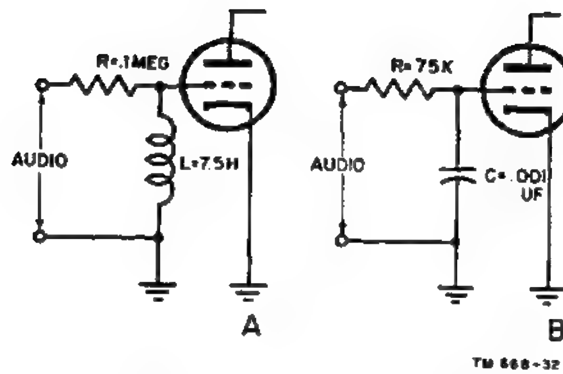


Figure 31. Preemphasis curve.



A. Preemphasis B. Deemphasis

Figure 32. Networks.

tance of 100,000 ohms (.1 meg) and an inductance of 7.5 henrys.

$$\text{Time constant} = \frac{L}{R} = \frac{7.5}{.1} = 75 \text{ microseconds}$$

For the specific ratio of inductance to resistance in figure 32, the graph of output voltage in respect to input voltage is shown in figure 31.

d. Deemphasis Networks. The deemphasis at the receiver must be the reverse of the preemphasis characteristic. This is accomplished by making the time constant of the resistor and capacitor in B of figure 32, equal to that of the preemphasis circuit. Since capacitive reactance decreases with increased frequency, the voltage across it decreases as the frequency rises. When the proper time constant is chosen, the higher frequencies are restored to their normal values. If the capacitor is in microfarads and the resistance is given in ohms, the product of R times C gives the time constant in microseconds. For example, in the circuit in B, the capacitor is .001 microfarad and the resistance is 75,000 ohms. What is the time constant?

$$\text{Time constant} = R \times C = 75{,}000 \times .001 = 75 \text{ microseconds}$$

This is the same time constant as that of the inductor and resistor in A, figure 32.

14. Noise and Communication

a. General. One of the greatest disadvantages in transmitting information by amplitude modulation is the susceptibility of an a-m signal to both natural and man-made noises. Some of the disadvantages of a-m with regard to noise and freedom from interference can be overcome by using frequency modulation. Although f-m is not the most efficient system for overcoming noise, it is one of the easiest methods to apply. Modulation by means of short pulses of energy is more efficient, but the transmitters are much more complicated than those used for f-m. They seldom are employed except where large, cumbersome equipment can be used.

b. Impulse and Fluctuation Noise. Most man-made noises fall into two general classifications; *impulse noise* and *fluctuation noise*. *Impulse noise* consists of sharp pulses of r-f voltage which, when detected in a receiver, take the form of equally sharp pulses of audio voltage. They are often many hundreds of times greater in amplitude than the desired signal and make it impossible for the desired signal to be received. Perhaps the most common producers of impulse noise are the ignition systems of gasoline engines. Since many radio applications call for the installation of communication equipment in military vehicles, impulse noise is a problem. Steps are taken to eliminate as much of this noise as possible in military vehicles, but such elimination measures can never be perfect. There is always a residual component of the noise which may cause serious difficulties if the received signals are weak. The second kind of man-made noise, called *fluctuation noise*, is of a more continuous character. It appears as a broad band of many pulses which bear little or no relation to each other. Such noises are produced to a great extent by rotating electrical machinery, gas rectifiers, high-voltage transmission lines, and similar power devices. The noise from a small motor, although frequently weaker than the signal by a considerable amount, is capable of causing severe interference and possible interruption to a-m reception.

c. Noise Transmission. Man-made noise can reach the receiver in several possible ways. It can be received as a radiated signal along with the desired signal, or it can be picked up at the input of the receiver by transfer of energy through the capacitance between the antenna and the noise-producing device. Power lines in the vicinity of the antenna to which a noise-producing device is connected may induce a

noise directly in the antenna, or the noise can be transmitted directly over the power line to the receiver itself.

d. Man-Made Noise Frequencies. Man-made noise is not distributed uniformly through the spectrum. Impulse noise is most bothersome in the frequencies approximately from 15 mc to 160 mc and fluctuation noise generally is more severe at lower frequencies, and with a maximum intensity reached at frequencies well below 20 mc. The noise is most severe at the frequency where the device producing it is large enough to act as a good antenna. For example, vehicles whose dimensions approach a half-wavelength in the vicinity of 30 mc produce their most severe ignition noise in this region. Most equipment used for vehicular communication, as well as some important fixed stations, operates in this range. The noise produced by the discharge of current through a gas, mercury-vapor rectifiers, and similar devices has a spectrum that extends up to extremely high frequencies with very high intensities. It often is impossible to remove the noise from these devices because it is impractical to shield them.

e. Natural Noises. Natural noises disturbing to radio communication arise from various sources and may be either impulse or fluctuation types. Perhaps the most familiar are the frequent and overlapping noise in pulses produced by lightning discharges. This noise originates not only in local storms but also in the tropical storm centers, from which it is propagated as a radio wave to many parts of the earth. The signal strength of noise produced by local storms decreases directly with increasing frequency, the frequencies above 40 mc being less subject to this interference than the lower frequencies. The storms are more intense in summer than in winter, and, since most of the static occurs in the warm temperatures, the lowest noise levels are found in the colder weather. There are also weaker noise impulses thought to be produced by sources not on the earth, such as the noise attributable to sun spots. Naturally, nothing can be done to suppress natural noises at their sources, whether they originate near the receiver or not. Therefore it is desirable to have some other means of overcoming them.

f. Receiver Noise. The limiting factor governing the sensitivity of most high-frequency receivers is the amount of noise inherent in the receiver. Although there may be no defective parts, there still will be a quantity of noise in even the best receivers. This noise is of a random fluctuation type, producing a characteristic hiss-like sound in the loudspeaker or earphones. The most important noise in very-high-frequency equipment, where noise outside the receiver is low enough to permit receiver noise to be noticed, is caused by the current flowing through resistors and tubes.

g. Communication in Presence of Noise. A general communication system is shown in the block diagram of figure 33. The information source supplies a message which is converted into an electrical impulse. It then is trans-

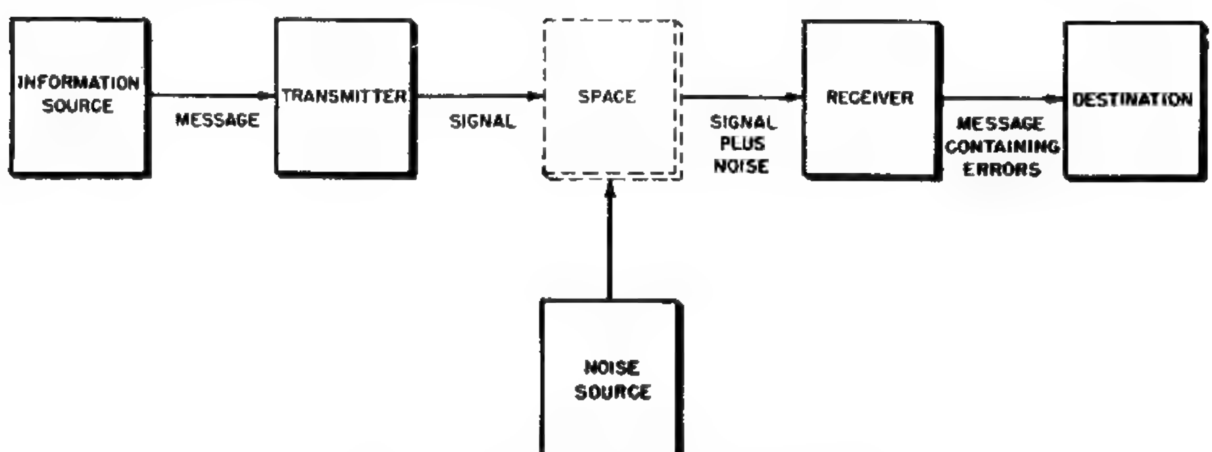


Figure 33. Noise in general communication system.

formed into a signal that can be sent through space to the receiver. The receiver recovers the signal and reproduces the message. During the course of the transmission, the signal may be modified by the addition of some noise, and therefore the receiver cannot reconstruct the original message perfectly. However, in an f-m system, if the deviation is increased above a certain minimum signal-to-noise ratio, the message-to-error (or degree-of-accuracy) ratio increases.

h. Noise in F-M Reception. The effect of random fluctuation noise and impulse noise in f-m receivers differs. Fluctuation noise pulses excite tuned circuits in the receiver, causing them to oscillate at their resonant frequency. The oscillations interfere with the carrier, causing spurious noise to appear in the detector output, and then diminish gradually until the next noise pulse. Impulse noise is largely a sudden disturbance in the amplitude of the signal. It is removed by the f-m detector, which does not respond to the amplitude variations of the carrier. If the frequency-modulated signal is very weak compared with the noise, the intelligence-containing side bands are suppressed in the f-m detector. Therefore, if the carrier is not above a certain minimum amount, f-m is actually worse than a-m. The value of this minimum is called the *threshold of improvement* and depends on the frequency deviation. Only when the signal is above this threshold is wide-band f-m superior to a-m. Narrow-band f-m is as good as a-m at low signal levels and is equal or inferior to it at high levels.

i. F-M and A-M Response to Impulse and Fluctuation Noise. In figure 34 the signal-to-noise ratio is plotted vertically against the peak carrier-to-noise ratio. Amplitude modulation is shown by the straight diagonal line for comparison purposes. For a modulation index of 1, the dashed lines indicate the improvement in signal-to-noise ratio over a-m. The lower of the pair, labeled F_1, is the improvement for fluctuation noise, and the higher, I_1, is the improvement for impulse noise. The other set of dotted lines shows the same situation for a modulation index of 4, with curve F_4 for fluctuation noise and I_4 for impulse noise. The threshold of improvement begins where the carrier is sufficiently

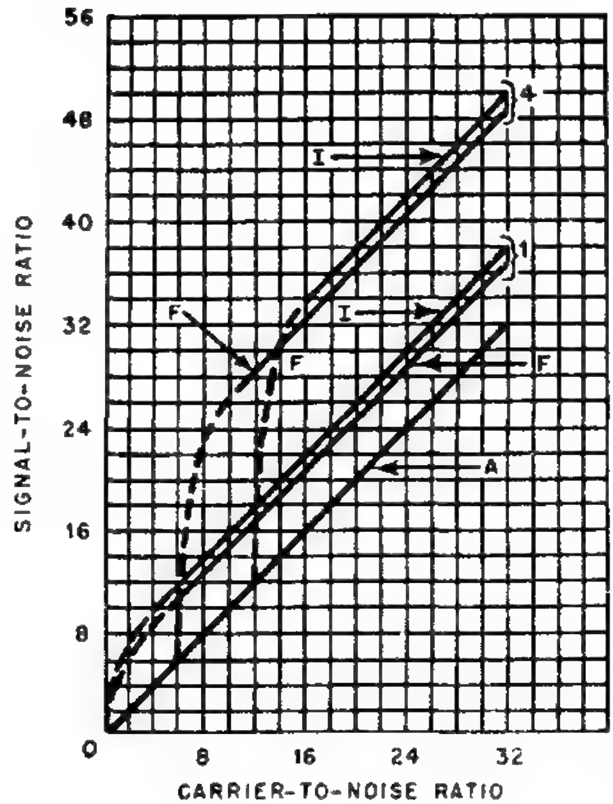


Figure 34. F-m noise characteristics.

strong that the curve crosses the reference a-m line.

j. Noise in Presence of Modulation. These analyses have been made with sine-wave modulation. With speech or other forms of information impressed on the carrier, the results are somewhat different, but in general they agree with that of the sine-wave. For normal military communication, f-m is an improvement over a-m because of its higher immunity to noise. *Narrow-band* f-m is preferred where the carrier frequency is relatively low and channel space is at a premium. *Wide-band* f-m is used at the higher frequencies, where this is not a problem. However, when the signal is weak, wide-band f-m is inferior in performance to an equivalent a-m system. In this it differs from narrow-band f-m, which is capable of excellent performance at low-signal levels. The reduction in errors caused by poor signal-to-noise ratio also makes advantageous the use of f-m in radiotelegraphy systems. In these systems, the sig-

nal is modulated by either a series of rectangular pulses or the use of two continuous audio tones as modulating signals to shift from one tone to the other as the transmitter is keyed. The latter method has the advantage of requiring somewhat less spectrum space than the rectangular modulation method.

15. Interference

a. General.

(1) Any receiver of radio signals, no matter what type of modulation is used, has a certain bandwidth over which it will accept signals. In this *acceptance band*, the receiver reproduces signals at the output of the detector for which it was designed. An a-m receiver must be capable of passing all frequencies that go into the modulation side bands on either side of the carrier in addition to the carrier itself. If speech is to be received, the receiver must have a bandwidth at least twice the highest speech frequency that is transmitted. For good voice transmission the highest audio frequency is usually 3 kc. Therefore, the bandwidth of an a-m receiver for the reception of speech must have an acceptance band of at least 6 kc.

(2) An f-m receiver must have a bandwidth that is at least enough to allow for the full frequency deviation plus the side bands of the transmitted signal. In a receiver with a given bandwidth, any spurious or unwanted signals that fall within the acceptance band give rise to some form of interference to the desired signal. In addition, no practical receiver is made with an acceptance band that just covers the necessary bandwidth with no response outside those limits. If the signal that interferes with the desired signal is in the same channel, the interference is called *co-channel interference*. If the disturbance originates on either side of the receiver acceptance band, it usually is called *adjacent channel interference*. The receiver also can have *spurious responses* at frequencies other than the one at which it is tuned because of inadequacies in the rejection of unwanted frequencies in the first receiver stage.

b. A-M.

(1) Within the acceptance band of the a-m receiver, any unwanted side bands are equivalent to spurious modulation, and they are amplified and detected just as the desired signal is. Therefore, no matter what steps are taken in the design of the receiver, these undesired signals cause an output to appear in the loudspeaker or earphones. The behavior of a receiver in the presence of interference can be analyzed in connection with the over-all acceptance band of the receiver, and the results for a-m are different from those for f-m.

(2) To see what happens when two signals combine in the pass band of any type of receiver, consider the vector diagram of figure 35. In this diagram, the undesired carrier is represented by the vector labeled E_i. It is not a stationary vector but is generating a sine wave and therefore is rotating at a definite rate depending on the frequency. Assume that it combines with the desired carrier, E_c, which has a frequency that is slightly different from that of the undesired carrier, E_i. Although both carriers are rotating, it is assumed that E_c is standing still while E_i rotates at a rate equal to the difference between its natural frequency and that of the desired carrier. Since the two carriers are added in the receiver, the resultant vector shown by the arrow, E_r, has a frequency equal to the difference between the two carriers. Moreover, its relative phase, a, is continuously changing with the rotation of the carriers, so that the resultant is phase-modulated. Since the frequency difference between the two signals was assumed to be small, the vector changes in length as the relative angles of its components

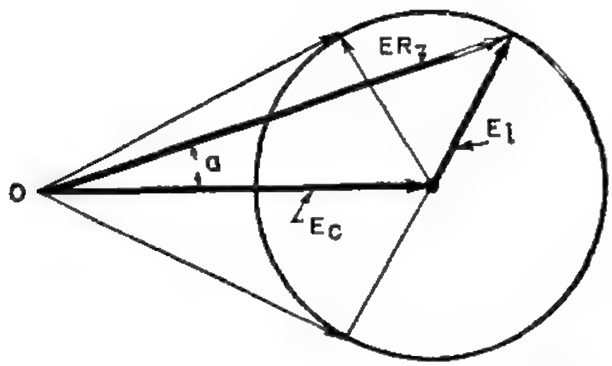


Figure 35. Interference between two unmodulated carriers.

vary. Therefore, it is also amplitude-modulated at the difference frequency.

(3) The single interference resultant produced between two unmodulated carriers is modulated in both amplitude and phase. Since frequency and phase modulation differ so little, it can be considered modulated also in frequency. This composite modulation is called a *beat note* between the two carriers. When this beat note is received, it produces spurious signals that vary both in frequency and in amplitude and therefore appear in the output of the receiver, regardless of the kind of modulation for which the receiver is designed. The output of the detector also contains simple multiples of the fundamental beat note. These are its harmonics and they are strong when the two carriers are nearly equal in amplitude.

(4) When one of the carriers is amplitude-modulated, the resultant interference varies at the modulating frequency. The beat note here too is frequency- , phase- , and amplitude-modulated, and can interfere in any receiver. If both carriers are amplitude-modulated, the resultant beat note is heavily amplitude-modulated, even though the strength of the interfering carrier is a fraction of the strength of the desired one. In fact, it can be shown by mathematical analysis that the interfering signal can be as much as one thousand times weaker than the desired carrier and still cause an audible beat note. If it is more than one-hundredth as strong, it causes prohibitive interference. Therefore, it can be seen that for military purposes, where there must be many pieces of equipment in use at once, interference-free reception with a-m may be impossible under emergency conditions.

c. F-M. In f-m, modulation of an otherwise unmodulated carrier by the undesired signal decreases as the deviation of the undesired carrier increases. In fact, it never can exceed one-half cycle, which is equivalent to a modulation index of one-half. When the desired carrier is modulated, the interference decreases as the deviation is increased, especially since the f-m detector is assumed to be insensitive to a-m. Wide-band f-m is less susceptible to interference than a-m, and narrow-band f-m gives some improvement. If the two interfering signals differ in amplitude by a relatively small ratio of two to one, the interference between the carriers is negligible. Therefore, because of its greater freedom from interference, f-m is more widely used for military applications than a-m. In f-m, the stronger of two wide-band, or even two narrow-band, transmitters is received practically without interference from the other.

16. Summary

a. In a frequency-modulated wave, the instantaneous frequency varies about the carrier frequency in proportion to the amplitude of the modulating signal.

b. The variations in instantaneous frequency are determined by the frequency of the modulating wave; the higher the modulating frequency, the greater the number of deviations in a given time period.

c. In amplitude modulation, two side-band frequencies (an upper and a lower) are generated for each sine-wave component of the modulating signal.

d. In an f-m system, many side bands are generated for each sine-wave component of the modulating signal.

e. The bandwidth occupied by an f-m signal exceeds the peak deviation limits. The total bandwidth increases for increasing modulation frequency, all other things being constant.

f. When the channel space occupied by an f-m transmitter approximates that of an a-m transmitter for the same modulating signal, the transmission is termed narrow-band f-m.

g. Where f-m bandwidth greatly exceeds that of the equivalent a-m signal, it is called wide-band f-m.

h. The amplitudes of the side bands, as well as their frequencies, depend on the amplitude and frequency of the modulating signal and the frequency deviation of the transmitter.

i. For a sinusoidally modulated signal, the side bands are distributed in symmetrical pairs on either side of the carrier frequency at integral multiples of the modulation frequency.

j. The ratio of the frequency deviation to the frequency of the modulating signal is called the modulation index.

k. The value of the modulation index determines the amplitude and number of the various side bands.

l. The number of side-band pairs increases as the modulating frequency decreases.

m. To allow for the side bands beyond the peak deviation limits, additional channel space, called a guard band, is provided on either side of the channel.

n. For nonsinusoidal modulation, the side-band frequency distribution depends on the sinusoidal components present in the nonsinusoidal wave, and can be analyzed in terms of each component separately plus all possible products of each side band with every other. The bandwidth for a nonsinusoidally modulated signal generally is greater than that for a sine wave of the same amplitude and frequency.

o. Where the modulation is in the form of rectangular waves, the result is termed pulse modulation. The number of pulses per second is called the pulse repetition rate.

p. Symmetrical modulating signals produce a symmetrical side-band spectrum.

q. Because the high frequencies in a sound signal are weaker than the low ones, they are amplified to a greater extent in many transmitters to improve the ratio of signal to noise in this range. This is called preemphasis. The reverse process is called deemphasis and is provided for in the receiver.

r. Radio noise is of two types—impulse, or sharp r-f pulse voltage; and continuous, or fluctuating, signals of random amplitude and phase.

s. Improvement in the signal-to-noise ratio takes place only with a minimum value of carrier present at a level termed the threshold of improvement.

t. Frequency modulation is less susceptible to interference than a-m because the maximum effective modulation index of the interfering wave can never exceed one-half if the desired carrier is twice as strong as the undesired carrier.

17. Review Questions

a. What happens to the frequency deviation of an f-m wave when the amplitude of the modulating signal is increased?

b. If the frequency of the modulating signal is doubled, what happens to the f-m wave? Describe in detail.

c. What determines the effective bandwidth of an a-m transmitter?

d. Why are guard bands used between adjacent f-m channels?

e. Are the side bands of an f-m transmitter the same for sinusoidal and nonsinusoidal modulating voltages?

f. How is it possible to use a larger bandwidth for an f-m transmission than for an a-m transmission?

g. When does f-m have only two side bands?

h. Which f-m signal has fewer side bands, one with a higher or one with a lower modulating frequency, for the same total deviation?

i. What is the minimum bandwidth of a narrow-band f-m signal?

j. If the audio-modulating frequency is the same for two f-m transmitters, but one is adjusted for one-half the deviation of the other, which will have the greatest effective number of side-band pairs? Why?

k. If a 5-kc audio signal produces a deviation of 15 kc at 33 mc, what is the maximum instantaneous frequency of the f-m wave?

l. What is meant by the modulation index?

m. What is the modulation index for an audio signal of 2 kc with a deviation of 220 kc? What total bandwidth would such a signal occupy?

n. Define pulse repetition rate.

o. What is the effect on the side-band spectrum when two signals of different amplitude and frequency simultaneously modulate the carrier?

p. When is the spectrum that results from nonsinusoidal modulation symmetrical? When is it not symmetrical?

q. What are the highest and the lowest frequencies that need to be transmitted for intelligible speech?

r. What is the difference in the relationship of phase-to-frequency deviation between a direct f-m transmitter and an indirect f-m transmitter, modulated by speech?

s. What is the purpose of preemphasis?

t. What is the time constant of a deemphasis network consisting of a capacitor of .003 microfarad in series with a resistor of .1 megohm?

u. What is the difference between impulse and fluctuation noise?

v. In an f-m communication system, what is the relationship between the bandwidth and noise?

w. By how much must two carriers differ in amplitude if they are not to interfere with each other when they are both amplitude-modulated?

x. Which is better for the reduction of interference for stations operating on the same channel—narrow- or wide-band f-m? Why?

CHAPTER 3

METHODS OF PRODUCING FREQUENCY MODULATION

Section I. DIRECT METHODS OF PRODUCING FREQUENCY MODULATION

18. General

a. Methods for producing f-m are *direct* and *indirect*. In the former, the change in frequency of the oscillator is directly proportional to the amplitude of the audio signal. With indirect methods, the change in phase angle is directly proportional to the amplitude of the audio signal.

b. The two basic transmitter stages necessary for producing f-m are the r-f oscillator and the modulator. The oscillator can be any one of the basic types, such as the Hartley or the Colpitts. The modulator stage provides a way of controlling oscillator frequency, usually with a modulating signal, at some audio frequency. A discussion of methods for producing f-m therefore is essentially a discussion of different modulator circuits.

c. Figure 36 shows a simple shunt-fed Hartley oscillator. Feedback is accomplished by coupling energy from the plate to the grid circuit through the split inductance. This feedback sustains oscillations at a resonant frequency determined mainly by the values of inductance and capacitance in the tank circuit. The frequency can be determined approximately by means of the formula,

$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

where

f_o is the resonant frequency
L is the inductance
C is the capacitance

From this formula, it can be seen that a change in either inductance or capacitance results in a change in the resonant frequency. If the inductance or capacitance is increased, the resonant frequency is lowered; if L or C is decreased, f_o is raised. Since the frequency can be changed by changing L or C, the next step is to find a method for allowing the audio signal to control either L or C.

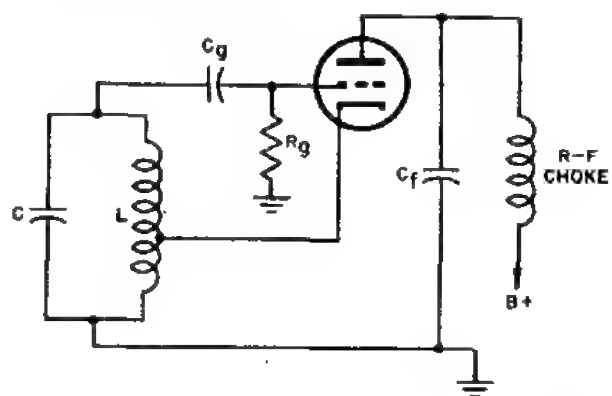


Figure 36. Shunt-fed Hartley oscillator.

19. Mechanical Methods

a. *Changing Capacitance*. In figure 37, a capacitor-type microphone is substituted for C in the tank circuit of the oscillator. The two plates of the microphone, one fixed and one movable, have some value of capacitance between them. This capacitance, C, together with the inductance, L, forms a tank circuit resonant at some frequency, f_o. When sound waves strike the movable plate (diaphragm) of the microphone, it moves back and forth with the changes in air pressure caused by the sound. From basic theory, it is known that changing the distance, d, between the plates of a capacitor changes its capacitance; the greater the distance, the

smaller the capacitance; the smaller the distance, the greater the capacitance. As the diaphragm moves back and forth, the capacitance of the tank circuit is being changed accordingly. A changing capacitance in the tank circuit means a changing resonant frequency. Since the audio signal controls the movement of the diaphragm and the value of capacitance, it also determines the resonant frequency. The output is a frequency-modulated wave, whose frequency deviation is controlled by the amplitude of the audio signal. There are certain drawbacks to using the circuit just described to produce an f-m wave since only a capacitor-type microphone can be used. This restriction is undesirable in view of the many types of microphones in use. Furthermore, it is necessary, at times, to separate the microphone from the transmitter by considerable distances. This arrangement is impossible with the circuit of figure 37.

specially designed signal generators. **However, the mechanical system necessary is not capable of changing frequency in accordance with voice-modulating frequencies. For this reason, it cannot be used in f-m transmitters.**

20. Reactance

Since the reactance of a capacitor, X_c, is equal to

$$X_c = \frac{1}{2\pi f_c} \text{ ohms}$$

Where

X_c is capacitive reactance

f is the frequency of the voltage source

C is the capacitance

It can be seen that, if the capacitance is held constant, the frequency must be changed in order to change the reactance. The same effect

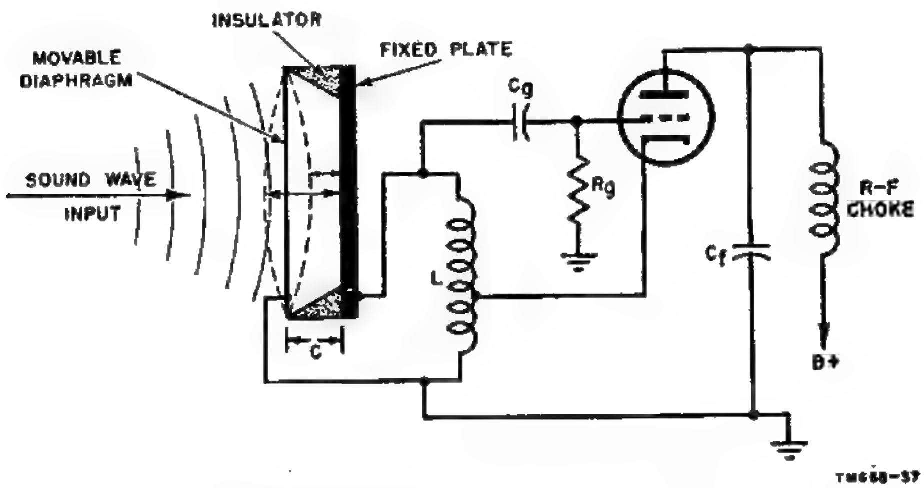


Figure 37. Capacitor-type microphone produces f-m.

b. Changing Inductance. The inductance of a coil constructed with a powdered-iron slug at its center can be varied by moving the slug in or out of the coil. If this coil is substituted for L in the tank circuit of the Hartley oscillator, the resonant frequency can be made to depend upon the position of the slug in the coil. By using a motor and a suitable gear train, the slug can be made to move in and out of the coil at a rate determined by the motor speed, and the resonant frequency is changed accordingly. This system is used to produce an f-m wave in

can be noted by inspecting the formula for inductive reactance:

$$X_L = 2\pi fL \text{ ohms,}$$

where

X_L is the inductive reactance

f is the frequency

L is the inductance

It is possible to control the frequency of an oscillator by controlling the amount of reactance, capacitive or inductive, which is present in the circuit. This is accomplished in a circuit known as a *reactance modulator,* which uses

the characteristics of a vacuum tube to control the reactance in the tank circuit of the oscillator. By simulating a capacitance or an inductance across its output terminals, it is said to *inject* reactance into the tank circuit. The simulated capacitance or inductance in turn is controlled by the audio signal.

21. Vacuum-Tube Characteristics

a. Transconductance. Before proceeding with the discussion of reactance modulators, it is desirable to review briefly the vacuum-tube characteristics on which circuit operation depends. The transconductance of a vacuum tube is defined as the ratio of a small change in plate current to the small change in the grid voltage that produced it, with the plate voltage held constant. This ratio can be expressed mathematically in the form:

$$g_m = \frac{di_b}{de_c}$$

where

g_m is the transconductance
i_b is the plate current
e_c is the grid voltage
d signifies *a small change in*

Transconductance is a form of conductance, and it is measured in *mhos*, where the current is in amperes and the voltage in volts. Because the transconductance almost never exceeds fractional values, a smaller unit, the *micromho*, one-millionth of a mho, is used. The transconductance of a vacuum tube under rated operating conditions usually is listed in tube manuals.

b. Amplification Factor. The amplification factor, μ, or mu, of a vacuum tube is defined as the ratio of a small change in plate voltage to the small change in grid voltage that produced it, the plate current being held constant. Expressed mathematically:

$$\mu = \frac{de_b}{de_c}$$

where

μ is the amplification factor
de_b is the change in plate voltage,
de_c is the change in grid voltage.

The μ of a tube determines how effective the grid is in controlling the plate current. For example, if the μ is 35, it means that the grid voltage is 35 times more effective in controlling the plate current than the plate voltage. For a given change in plate current, therefore, the change in plate voltage would have to be 35 times the necessary change in grid voltage.

c. Plate Resistance. The plate resistance of a vacuum tube is defined as the ratio of a small change in plate voltage to the small change in plate current that produced it, the grid voltage being held constant. In mathematical form—

$$r_p = \frac{de_b}{di_b}$$

where

r_p is the a-c resistance from plate to cathode
de_b is a small change in plate voltage
di_b is a small change in plate current

d. Relationship of g_m, μ, and r_p. It is possible to combine the three equations for transconductance, amplification factor, and plate resistance so that a direct relationship exists between them. If the equations for g_m and r_p are multiplied

$$g_m \times r_p = \frac{di_b}{de_c} \times \frac{de_b}{di_b}$$

The di_b terms cancel out, and

$$g_m \times r_p = \frac{de_b}{de_c}$$

An inspection of the equation shows that the right-hand side is the expression for the amplification factor, μ. Therefore, μ can be substituted for de_b/de_c, and

$$g_m \times r_p = \mu$$

The significance of this relationship between vacuum-tube characteristics will become apparent in considering reactance modulator circuits as used in f-m transmitters.

22. Reactance-Tube Modulator

The purpose of the reactance-tube modulator in a direct f-m transmitter is to frequency-modulate the f-m signal of an oscillator in accordance with the audio signal. The modulator accomplishes this by changing the amplitude variations of the audio signal into a varying reactance which is injected into the tank circuit of the oscillator. This injected reactance changes the frequency of the oscillator.

a. Basic Circuit.

(1) The fundamental arrangement of the circuit of a reactance-tube modulator and oscillator is shown in figure 38. At the extreme left, the audio signal is impressed between terminals 1 and

2, or between grid and ground of the reactance tube. Therefore, the audio signal controls the reactance-tube plate current. The plate current flows through a load consisting of Z_a and Z_b in series, which is connected across the tank circuit of the oscillator through terminals 3 and 4. When the plate current is varied by the audio signal, the reactance of the plate load is changed, thus changing the operating frequency of the oscillator tank circuit. The resultant f-m signal then is coupled inductively to the following stages of the transmitter through terminals 5 and 6.

(2) With no audio signal present at the grid of the reactance tube, its plate current has some small d-c value whose effect on the plate load is negligible. The r-f voltage present across the oscillator tank circuit also is impressed across the reactance tube plate load. The plate load of Z_a and Z_b presents a reactance to this voltage which can simulate the action of either a capacitor or an inductor. This is the same as adding a capacitor or an inductor in parallel with the tank circuit, which changes either the effective capacitance or the inductance of the tank circuit. The operating frequency of the oscillator, therefore, depends not only on L and C, but also on Z_a and Z_b. Since this operating frequency is the oscillator output frequency with no audio signal present, it also must be the center frequency of the f-m signal. The audio signal, when applied between terminals 1 and 2, is impressed not only from grid to ground, but also across Z_b in the plate circuit. The combined effect of these two actions is to change the reactance of Z_a and Z_b in accordance with the audio signal.

b. Impedance at Terminals 3 and 4.

(1) Looking from the tank circuit toward terminals 3 and 4, an impedance caused by Z_a, Z_b, and the reactance tube is seen. For a given amplitude of audio signal, the resultant plate current

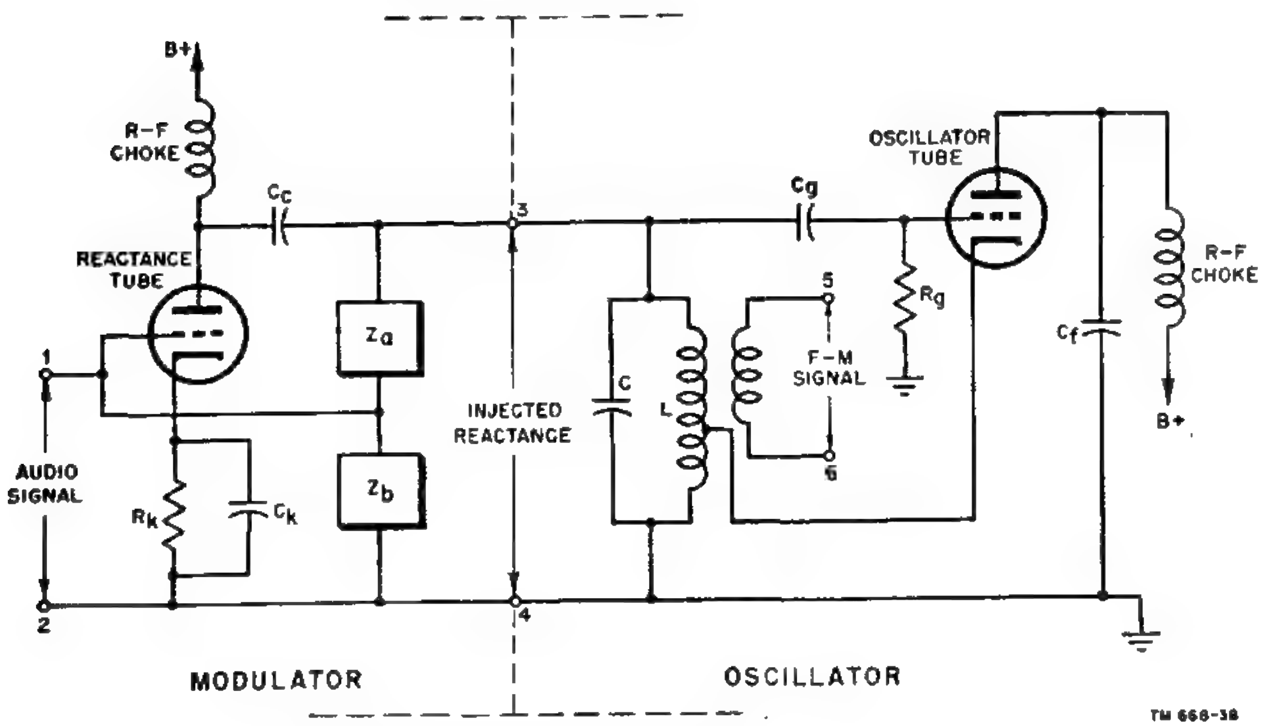


Figure 38. Basic circuit of reactance-tube modulator and oscillator.

through the tube can be found from from the tube transconductance

$$g_m = \frac{di_b}{de_c}$$

or

$$di_b = g_m \times de_c$$

The plate current flowing through the plate load is

$$i_b = g_m \times e_c$$

where the plate load consists of Z_a and Z_b, or Z_{ab}. In circuits of this type, the plate resistance of the tube usually is high enough to be ignored when considering the plate load.

(2) If the voltage across Z_{ab} is E_{ab}, then (by Ohm's law):

$$Z_{ab} = \frac{E_{ab}}{i_b}$$

Substituting for i_b,

$$Z_{ab} = \frac{E_{ab}}{g_m \times e_c}$$

From this it can be seen that the transconductance of the tube has an inverse effect on impedance Z_{ab} across the oscillator tank circuit. It remains now to break e_c into its equivalent components in order to observe all the factors at work in the determination of this impedance.

(3) An examination of figure 38 shows that e_c is impressed across Z_b in the plate circuit of the reactance tube. If the total voltage across Z_{ab} is E_{ab}, then e_c must be related to E_{ab} as Z_b is related to Z_{ab}. In mathematical form, this relation can be expressed as:

$$\frac{e_c}{E_{ab}} = \frac{Z_b}{Z_a + Z_b}$$

When this equation is solved for e_c, it is found that

$$e_c = E_{ab} \times \frac{Z_b}{Z_a + Z_b}$$

This value of e_c then is substituted in the equation for the total impedance found in the previous paragraph:

$$Z_{ab} = \frac{E_{ab}}{g_m \times e_c}$$

$$Z_{ab} = \frac{E_{ab}}{g_m \times E_{ab} \times \frac{Z_b}{Z_a + Z_b}}$$

The E_{ab} terms cancel out, and the remaining terms of the equation can be simplified in the following manner:

$$Z_{ab} = \frac{1}{g_m} \times \frac{Z_a + Z_b}{Z_b}$$

$$= \frac{1}{g_m} \times \frac{(1 + Z_a)}{Z_b}$$

$$= \frac{1}{g_m} + \frac{1}{g_m} \times \frac{Z_a}{Z_b}$$

The last form of the equation gives the impedance at terminals 3 and 4 in terms of the reactance-tube transconductance and the impedance network of Z_a and Z_b.

(4) This equation for the impedance seen by the oscillator tank circuit expresses mathematically the principle of operation of the reactance-tube modulator. The impedance, Z_{ab}, is the general term for the total impedance across the oscillator tank circuit. The terms Z_a and Z_b are expressions for the circuit components which together constitute the plate load. These can be capacitors, inductors, or resistors which have *fixed values* in the circuit, and the only way to vary Z_{ab} is to vary the transconductance of the reactance tube. This takes place when an audio signal is applied to the input to the modulator.

c. Injected Reactance. An examination of the equation for the total impedance shows that it contains two parts. The first part is the reciprocal of a transconductance, which is resistance. The second part contains two terms for impedance, Z_a and Z_b. If either Z_a *or* Z_b is made reactive, the second part of the equation, $1/g_m$ times Z_a/Z_b, is a form of reactance. This reactive component is the injected reactance. Assume that Z_a is a fixed capacitor having a reactance of X_c at a frequency, *f*. If Z_b is a resistor having a resistance, *R*, then the injected reactance, X_i, is

$$X_i = \frac{1}{g_m} \times \frac{X_c}{R}$$

d. Circuit Operation. Figure 39 shows a reactance-tube modulator circuit with C_L and R_L forming the plate load. With no audio signal present, this network injects a fixed reactance across the tank circuit which determines the

operating frequency of the oscillator or center frequency of the f-m wave. When an audio signal appears at the input, the injected reactance is varied above and below its zero-signal value, thereby varying the frequency of the oscillator above and below the center frequency. The variation in injected reactance, as well as the frequency deviation, depends on the amplitude of the modulating signal.

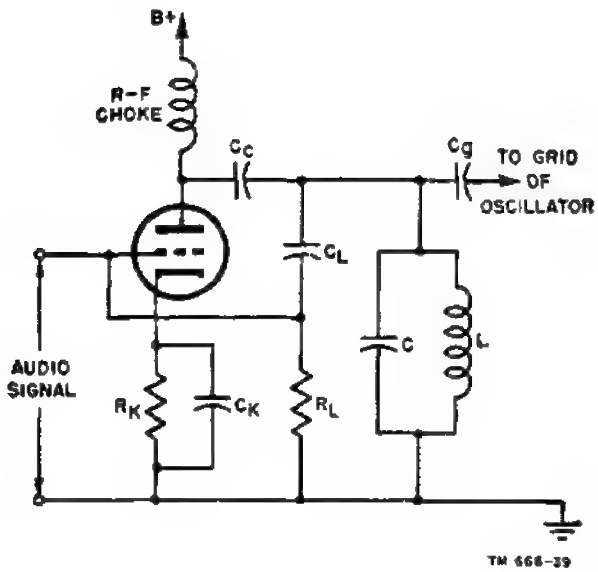


Figure 39. Typical reactance-tube modulator circuit.

(1) *Zero-signal operation.*

(*a*) With no audio signal present, the only signal impressed on the modulator circuit is that of the r-f voltage across the oscillator grid circuit. This voltage is applied across the plate load of the modulator, or across C_L and R_L. The reactance of C_L is made very large in respect to the resistance of R_L, and therefore the capacitive reactance, XC_L, determines the resultant current flow, causing it to lead the voltage across it by approximately 90°. This current flowing through R_L results in a voltage drop across R_L which also leads the applied voltage by 90°. Actually, the voltage across the resistance is in phase with the current through it; however, it can be seen that the voltage across R_L is applied between grid and ground of the reactance tube.

(*b*) This r-f voltage at the grid of the reactance tube causes an r-f variation of plate current which is coupled back to the tank circuit of the oscillator through the coupling capacitor, C_c. However, the current in the oscillator tank circuit is in phase with the r-f voltage, since the circuit is operating at resonance, while this additional current resulting from the *same* voltage leads the voltage by 90°. The additional current supplied by the reactance tube *acts as if it were caused by a capacitor.*

(*c*) The circuit operation just described can be seen in terms of vectors in figure 40. In A, the tank circuit voltage, E_T, and the current, I_T, are in phase because the circuit is at resonance. The r-f voltage of the tank circuit, E_T, causes a current flow through the plate load of the reactance tube. This current, i_L, leads the applied voltage by 90°, because the capacitive reactance of C_L has been made large in comparison with the resistance of R_L. In B, this current, i_L, flows through R_L, causing a voltage drop, E_{RL}, across the resistor which is in phase with the current through it. The circuit, however, is arranged so that this voltage is coupled to the grid of the reactance tube; E_{RL} is also e_c. In C, this grid voltage causes an r-f variation in plate current, i_b, which is in phase with the grid voltage. The capacitor, C_c, has negligible reactance and when the r-f current is coupled to the oscillator tank, it adds to the current flowing in it. Since i_b is 90° out of phase with the tank current and leading it, i_b acts as if it were coming from a circuit component having capacitive reactance, and consequently a capacitance.

(2) *Zero-signal injected reactance.* The amount of reactance injected in this

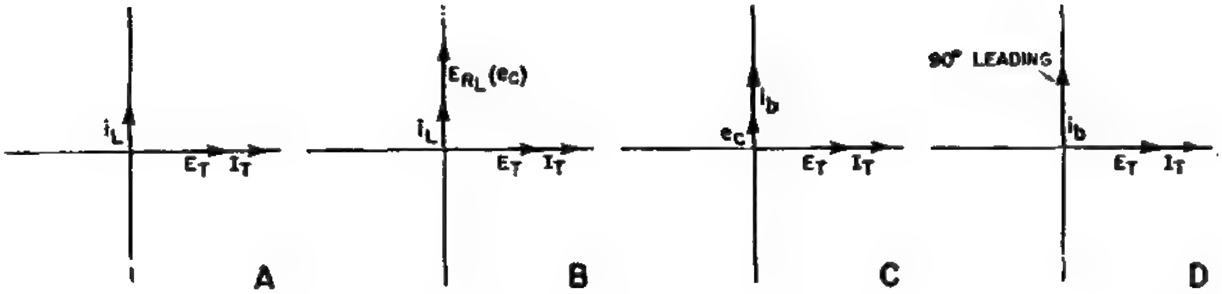


Figure 40. Vector representation of zero-signal operation.

circuit can be calculated from the formula

$$X_i = \frac{1}{g_m} \times \frac{X_{CL}}{R_L}$$

where

X_i is the injected reactance
g_m is the reactance-tube transconductance
X_{CL} is the capacitive reactance
R_L is the resistance

From the analysis of this particular arrangement of the plate load components, it is known that X_i is a capacitive reactance. Basic theory shows that the capacitive reactance of any capacitor can be found from the formula

$$X_c = \frac{1}{2\pi fC}$$

Since this is true, similar expressions can be written for the capacitive reactance of C_L and also for C_i,

$$X_c = \frac{1}{2\pi fC_L}$$

and

$$X_i = \frac{1}{2\pi fC_i}$$

Substituting these expressions for X_{CL} and X_i in the formula for the injected reactance:

$$\frac{1}{2\pi fC_i} = \frac{\frac{1}{2\pi fC_L}}{R} \times \frac{1}{g_m}$$

When this expression is simplified, the injected capacitance, C_i, is found to be

$$C_i = g_m \times R_L \times C_L$$

The effect of the reactance modulator, with no audio signal present, is the same as that of a capacitor having C_i capacitance placed across the oscillator tank. Since C_i is in parallel with C of the tank circuit, the two capacitances add, increasing the total capacitance and decreasing the operating frequency some fixed amount. The new operating frequency then becomes the center frequency of the f-m wave.

(3) *Effect of audio signal.* Applying an audio signal at some single frequency to the grid of the reactance tube causes two voltages to be present at the grid—an a-f voltage and an r-f voltage. The r-f voltage is responsible for the reactive plate current flow and the audio signal *changes* the amount of plate current flowing in accordance with its amplitude. Changing the amount of plate current coupled to the tank circuit means that its reactive effect is varied and results in the injection of a changing reactance into the oscillator tank. This changing reactance adds a changing capacitance to the oscillator tank. Oscillator frequency is varied accordingly and the result is a frequency-modulated signal at the oscillator output. The effect of the audio signal can be seen in terms of the formula for injected capacitance, where C_i is equal to $g_m R_L C_L$. The variations in a-f voltage at the grid have the same effect on the r-f plate current as a varying tube transconductance. From the formula above, it can be seen that varying the transconductance changes the injected capacitance, and causes frequency modulation to appear at the oscillator output.

e. Plate-Load Arrangements.

(1) The arrangement shown in figure 39 is not the only possible arrangement that can be used to inject a reactance into the oscillator. For example, the capacitor and resistor used in the basic circuit can be reversed so that *inductive* reactance is injected into the oscillator. A combination of an inductor and a resistor or any combination but that of capacitor and inductor allows the circuit to frequency-modulate the oscillator output. Four arrangements are possible: capacitor-resistor, resistor-capacitor, inductor-resistor, and resistor-inductor. However, the formula for the injected reactance involving g_m, Z_a, *and* Z_b has a different arrangement for each circuit.

(2) In figure 41, a resistor is substituted for Z_a and a capacitor for Z_b. The components are chosen so that the resistance of R_L is much greater than the reactance of C_L. Since the resistive component is so much larger, the r-f voltage applied to the plate load by the tank circuit causes the current to be in phase with the r-f voltage. As in any capacitor, however, the current leads the voltage by 90°, and this is true also of C_L. The voltage across C_L, therefore, lags the current and the applied voltage by 90°. This voltage across C_L is coupled to the grid of the reactance tube, and causes an r-f variation in plate current that is in phase with the grid voltage. The r-f current is coupled to the oscillator tank, and, since it is in phase with the grid voltage, it must lag the current in the tank by 90°. This produces the same result as when an *inductor* is placed across the tank. The modulator therefore is said to inject inductive reactance into the tank circuit. The amount of simulated inductance necessary to produce this injected inductive reactance is:

$$L_i = \frac{R_L C_L}{g_m}$$

(3) Keeping the large resistor, R_L, for Z_a, it is possible to substitute a small in-

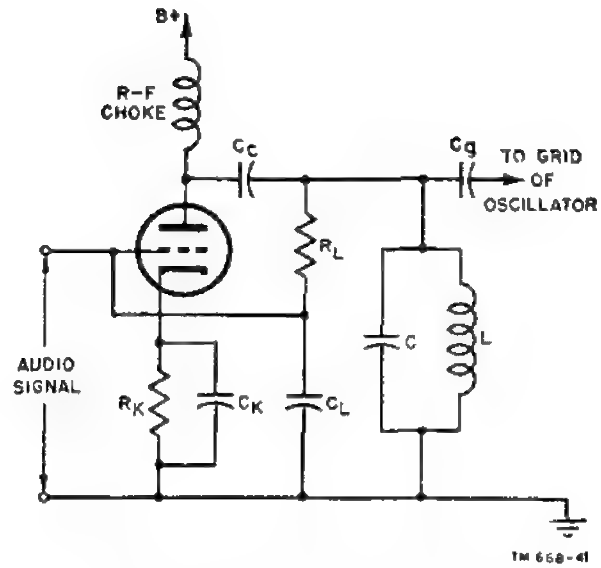


Figure 41. R-c reactance-tube load.

ductor for Z_b (fig. 42). With this arrangement, the circuit now injects a *capacitive* reactance into the oscillator tank circuit. The oscillator voltage applied across the plate load of the reactance tube causes a current to flow whose phase is controlled by the large resistance of R_L. This current is in phase with the applied voltage, since R_L is large in respect to X_{LL}. The voltage across any inductor, however, leads the current through it by 90°. The voltage across the inductor, therefore, leads both the current and the applied voltage by that amount. Because this voltage is coupled to the grid of the reactance tube, an r-f plate current flows which is in phase with the grid voltage and 90° leading in respect to the oscillator tank voltage. When coupled to the oscillator tank, this current acts as if it were caused by a *capacitor* having a capacitive reactance equal to the injected reactance. The amount of this capacitance can be found from the formula,

$$C_i = \frac{g_m L_L}{R_L}$$

(4) When the resistor-inductor arrangement is reversed (fig. 43), the circuit injects *inductive* reactance into the oscillator tank. The inductance of L_L is

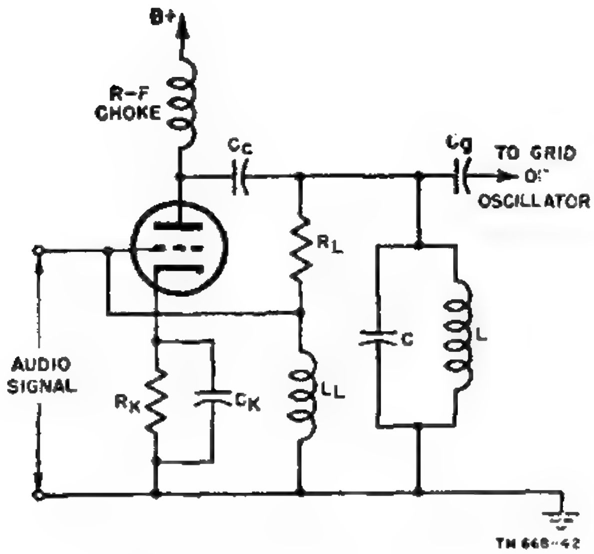


Figure 42. R-L reactance-tube load.

made large to increase the inductive reactance in respect to R_L. The r-f voltage from the oscillator tank then will cause a current to flow through the plate load which lags the applied voltage by 90°. This voltage then is applied to the grid of the reactance tube, producing an r-f plate current which is lagging the current in the tank circuit by 90°. Since a lagging current is an indication of inductance, the modulator simulates the action of an inductor placed across the tank circuit. The amount of injected inductance can be found from the formula

$$L_i = \frac{L_L}{g_m R_L}$$

(5) The four arrangements possible in the plate circuit of the reactance tube are shown in figure 44, together with the formulas for calculating the injected inductance or capacitance. It must be borne in mind that Z_a always is made large in respect to Z_b, and it is this *component which determines the phase* of the current caused by the r-f oscillator voltage. The amplitude of the reactive current is controlled by the amplitude of the audio-modulating signal by means of the tube transconductance. Z_a and Z_b, therefore, control the operating frequency of the oscillator, and g_m controls the frequency deviation.

f. Quadrature. The basic reactance-tube circuit *often is referred to as a quadrature circuit* because the r-f voltage developed across its output terminals is leading or lagging the r-f current in its plate circuit by approximately 90°. When the injected reactance is capacitive, the current leads the r-f voltage; when it is inductive, it lags. If the plate resistance of the reactance tube is negligible in respect to the magnitude of the injected reactance, the phase angle approaches the value of 90°. However, since there is always a resistive component as well as a reactive one, this phase angle can never actually become the ideal quadrature relation of 90°.

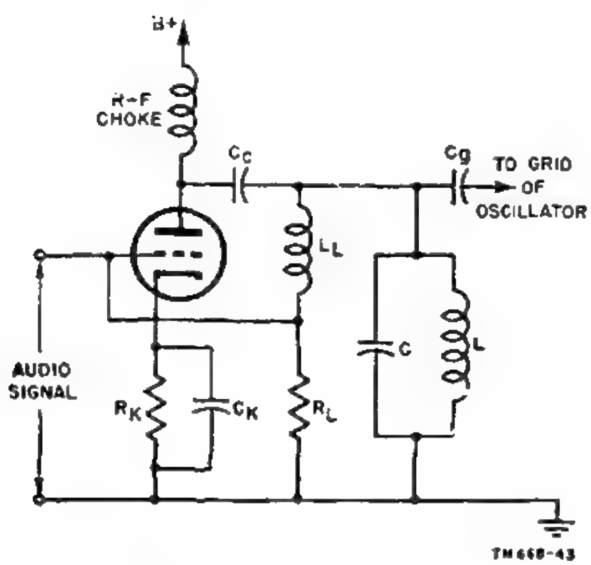


Figure 43. L-R reactance-tube load.

23. Practical Reactance Modulator

A of figure 45, shows a reactance modulator circuit used in a portable f-m transmitter, which injects a capacitance across the tank circuit of the master oscillator. For explanation purposes, the equivalent circuit of A is shown in B. The plate tank circuit of the master oscillator is represented by L_1-C_1 and the interelectrode capacitance between the grid and filament of the reactance modulator by C_{gf}. The interelectrode capacitance, C_{gf}, is connected effectively in series with $R75$ and in parallel with $L2$ to make up a phase-shifting network. Since

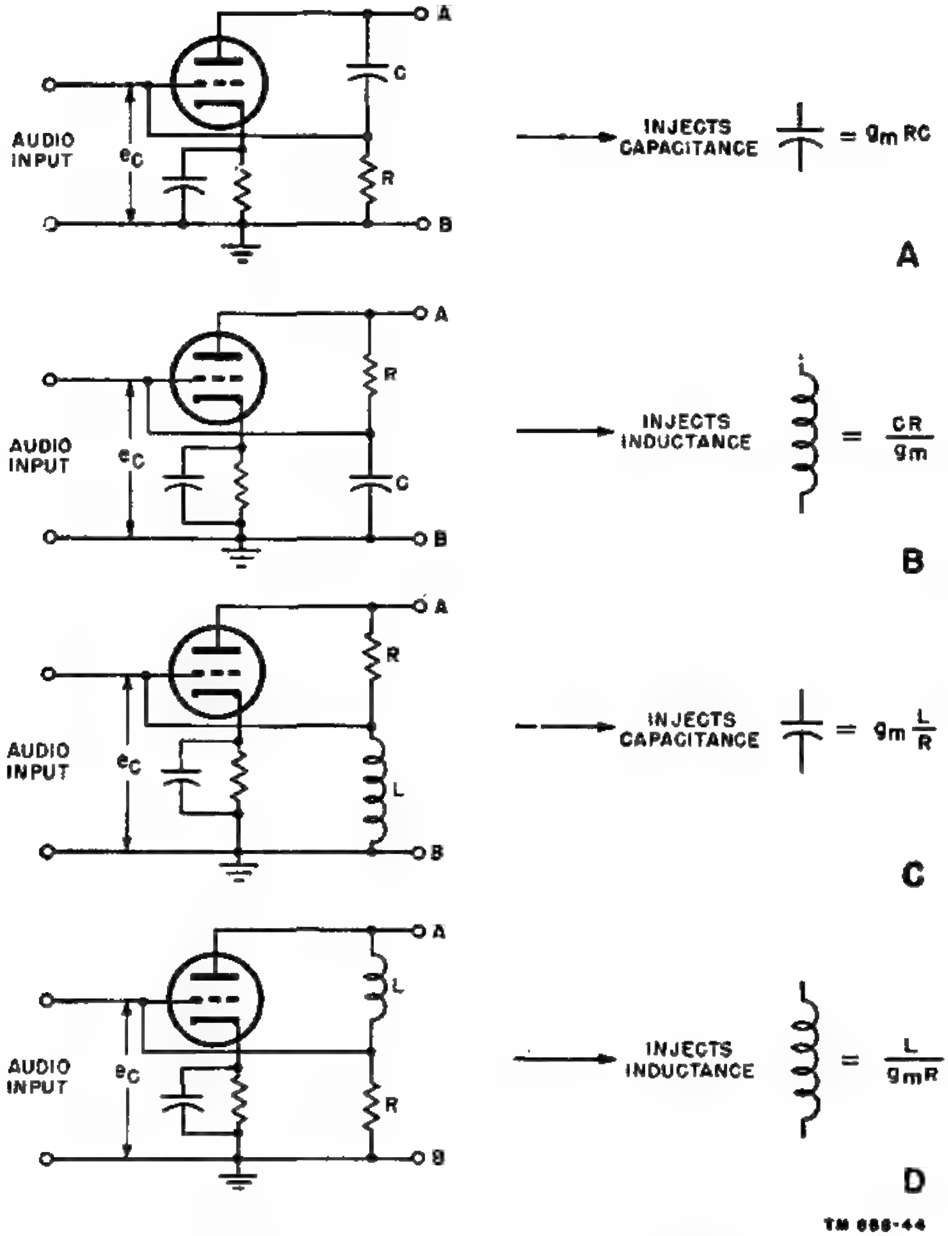


Figure 44. Four possible plate-load arrangements.

L_2 is coupled to L_1, the r-f voltage across it is 180° out-of-phase with the oscillator voltage across the tank. The resistance of *R*75 is made large in respect to the reactance of C_{gf}, and the current through C_{gf} is in phase with the induced voltage produced by *L*2. However, the voltage across C_{gf} lags the current through it by 90°, and this voltage is applied to the grid of the reactance tube. Since the grid voltage is in phase with the plate current, it lags the tank voltage and current by 180° plus 90°, or 270°. This is the same as saying that the plate-current leads the tank current by 90°. The effect is that of injecting a capacitance across *L*1-*C*1. Capacitor *C*28 blocks the d-c bias voltage on the oscillator grid from appearing on the reactance modulator grid. The purpose of *R*17 is similar to that of an r-f choke; it keeps the low-impedance microphone circuit from shorting the r-f components across C_{gf}.

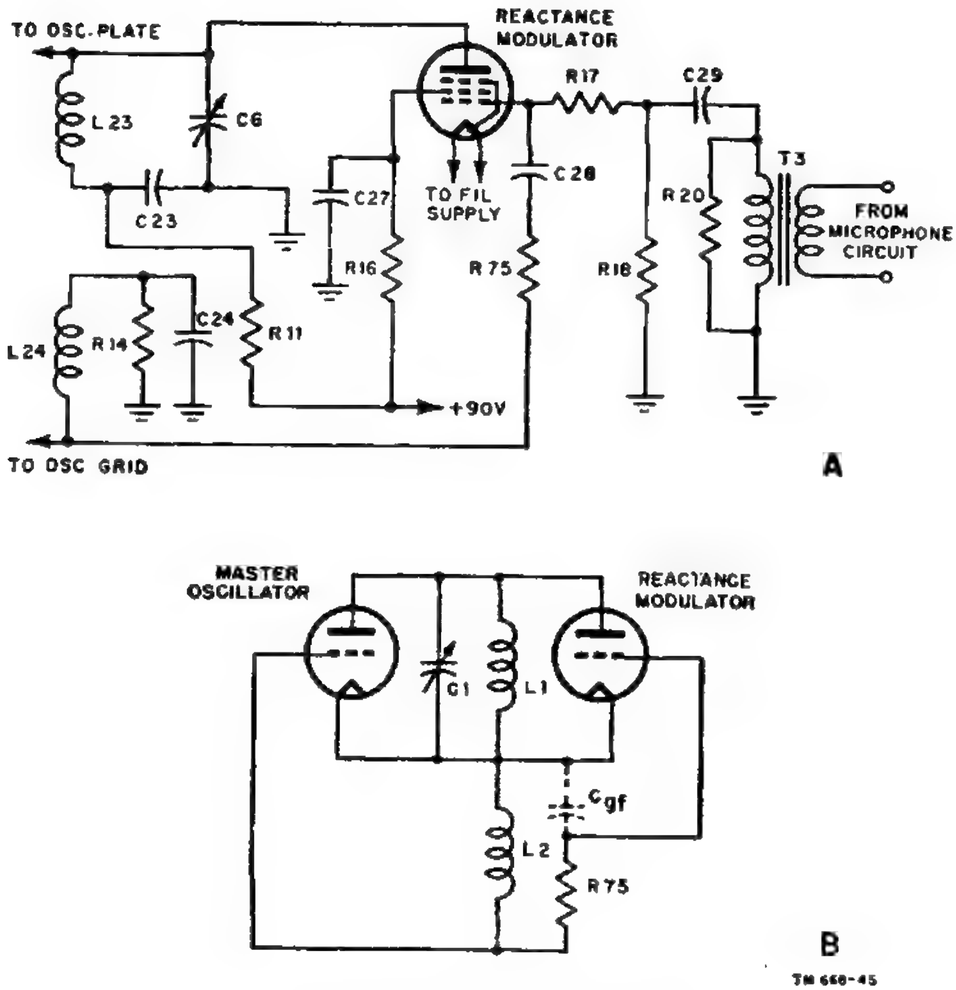


Figure 45. Practical reactance modulator.

24. Input Capacitance Modulator

a. In any vacuum-tube amplifier a definite impedance is present between the grid and the cathode. This impedance depends on the values of components in the external circuits of the other tube elements. Resistance in the plate circuit will result in the appearance of capacitance from grid to ground, depending on the amount of series plate resistance and the voltage gain of the tube. This is caused by the variations in electrical charge in the space between the grid and the plate. In the tube manual, a capacitance from grid to ground usually is listed for each tube under the heading of *input* capacitance. This value is determined with no resistance in the plate circuit. As the plate load is increased in value, the capacitance reflected into the grid circuit rises. How effective a given amount of plate load is in increasing this capacitance depends on the gain of the tube, and ultimately on the transconductance, $\mu = r_p g_m$. The changing of input capacitance with variations in plate-load resistance and tube transconductance is called *Miller effect.*

b. Since the input capacitance varies with the transconductance, it can change the frequency of an r-f oscillator when connected across the tank circuit. Therefore, Miller effect can be put to work in a direct frequency modulator (fig. 46). Operation is much like that of the reactance modulator, since variations in grid voltage at an audio rate are reproduced as changes in frequency of the oscillator. With no audio signal on the modulator, a fixed capacitance is injected across the oscillator tank. Consequently, the effective frequency of oscillation

goes down to the value set by the sum of the Miller-effect capacitance and the oscillator tank capacitance. When the audio signal is applied, the transconductance of the amplifier tube changes directly with the amplitude of that signal. The value of the Miller-effect capacitance increases and decreases with the amplitude variations of the audio signal. Since this capacitance is in parallel with the oscillator tank capacitor, the frequency varies above and below the audio-signal zero value, producing true f-m.

for wide deviations as the reactance modulator. Furthermore, the amount of injected capacitance is not adjustable over the wide range that the quadrature circuit offers.

25. Diode Modulator

a. When a resistor and a capacitor are placed in series across a source of a-c voltage, the current flowing in the circuit is out of phase with the voltage. This current has two components, one caused by the resistance and one by the ca-

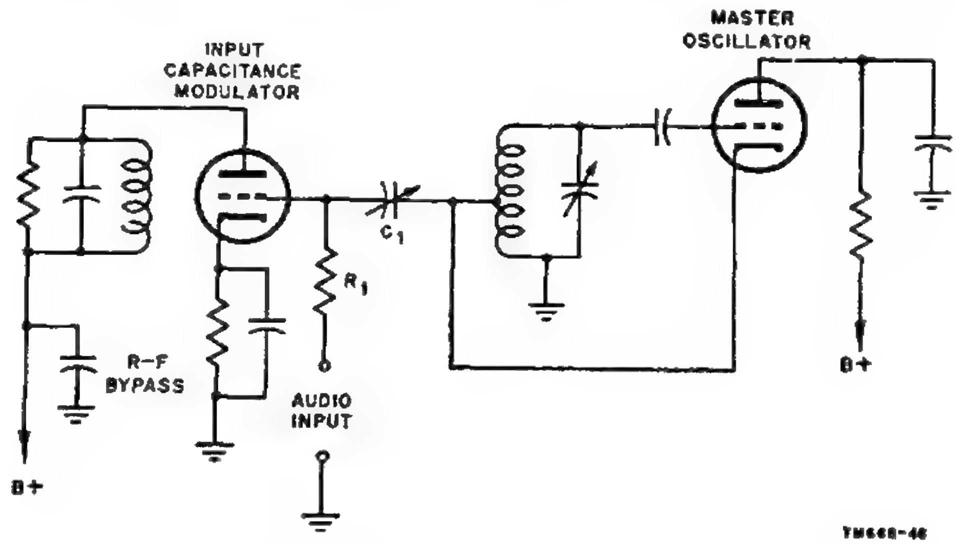


Figure 46. Input-capacitance modulator.

c. The amount of capacitance injected by Miller effect also depends on the internal tube capacitance between grid and plate. In a pentode, this internal capacitance is small, and the Miller effect is also small; in a triode, the internal tube capacitance itself usually is sufficient. To increase the amount of injected capacitance, a small capacitor can be connected externally between the grid and plate. A small value of plate load resistance is used to permit the tube to operate along a linear part of the plate current-grid voltage characteristic curve. Cathode bias is provided by a series resistor and bypass capacitor. To keep the deviation distortion low, the tube is coupled through *C*1 to a tap across a part of the oscillator tank. Audio voltage is applied to the grid of the tube through an isolating resistor, *R*1. The circuit is capable of a considerable amount of frequency deviation with low distortion. However, it is not as good

pacitive reactance. The resistive component is in phase with the applied voltage, and the reactive component leads the applied voltage by 90°. The resultant current therefore leads the applied voltage by some angle between 0° and 90°, the size of the angle depending on the relative size of the two components. If the resistance is large compared with the capacitive reactance, then the current leads the applied voltage by only a few degrees. If the resistance is small compared with the capacitive reactance, the current leads by nearly 90°.

b. If the resistance is made variable, the current can be made to lead the voltage by any value between 0° and 90°. If this variation can be controlled at the audio rate of a modulating signal, the resultant current can be used to frequency-modulate an oscillator. A diode is used as the variable resistance in the arrangement shown in figure 47, called a *diode modulator.*

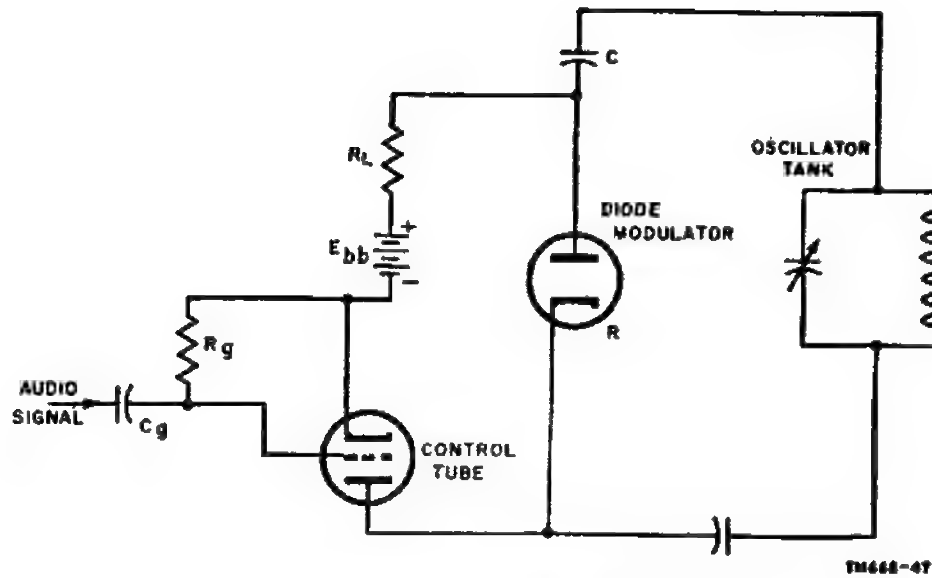


Figure 47. Diode modulator.

c. The diode and capacitor C are placed in series across the oscillator tank circuit. The reactance of C is made large in respect to the resistance of the diode. The applied voltage, E is the r-f voltage across the tank, and causes an r-f current to flow through the diode and C, which leads the voltage by nearly 90°. The plate current of the control tube also flows through the diode and R_L to the power supply. With zero-modulating signal this current is constant, and the modulator injects a fixed reactive current into the tank which determines the operating frequency.

d. When an audio-modulating signal is applied to the grid through C_g, an audio current flows through the plate circuit of the control tube. This current increases and decreases in step with the audio voltage. Since the current through the control tube also flows through the diode, the current flow through the diode also increases and decreases. This has the same effect as increasing and decreasing the resistance of the diode with a constant voltage applied. The *effective* resistance of the diode in series with the r-f current, therefore, is varied at the audio rate. The variations in the resultant reactive current injected into the tank circuit change the frequency of oscillation. A change in the reactive current has the same effect as a change in the capacitance across the tank circuit, and the result is a frequency-modulated signal.

e. Less noise and distortion are present in the diode modulator than in the conventional reactance-tube modulator. In addition, it is easy to set the operating range so that the frequency deviation is more nearly proportional to the modulating signal. The reactance-tube modulator has considerable variation in the resistive current drawn during modulation which always introduces some amplitude modulation. In the diode modulator, the actual resistive change that takes place across the oscillator tank is so small that there is very little resultant distortion. The reactance modulator, however, is capable of a much greater maximum frequency deviation.

26. Frequency Modulation of R-C Oscillator

a. Several communication systems have been discussed in which the information transmitted was not necessarily speech. One of the most common of the nonspeech applications of f-m is facsimile, involving the transmission of pictures, maps, and similar material which are printed automatically at the receiver in response to signals from the transmitter. Facsimile often uses a system of f-m known as *sub-*

carrier modulation which consists of the amplitude modulation of an r-f carrier by an audio tone. The a-m wave then is frequency-modulated and serves as a carrier for the applied signal.

b. Subcarrier modulation requires wide deviation relative to the carrier frequency. For example, if the subcarrier is 4,000 cps, it is common to find deviation which varies the frequency between the limits of 2,000 and 6,000 cps. This variation is more than 50 percent of the subcarrier frequency itself. Such relatively wide deviation cannot be obtained from a reactance modulator driving an ordinary *L-C* oscillator without excessive distortion. The variation of transconductance necessary to produce the relatively enormous change in injected capacitance or inductance needed cannot be obtained from the straight-line part of the e_c-i_b characteristic. Therefore, a completely different solution has been worked out, using an oscillator that has no inductance in its circuit.

c. Oscillators for the audio-frequency range can be made using only resistors and capacitors. Their operation is based on the principle that a series combination of a resistor and a capacitor causes a definite phase shift. When the values of the resistance and the capacitive reactances coincide, the current in the circuit leads the applied voltage by 45°. If the voltage developed by the current in the resistor is applied to a second identical series network, the current in this circuit leads the original input circuit voltage by 90°. If the current is made to flow through two more identical networks (fig. 48), the overall phase shift from the input to the output of the network (4 resistors and 4 capacitors) is 180° at a particular frequency.

d. The plate voltage of a vacuum tube is opposite in polarity to that of the voltage on the grid, and these voltages can be considered 180° out of phase with each other, when the plate load is resistive. A decrease in grid voltage causes a corresponding decrease in plate current. If this decreased plate current flows through a load resistance, the voltage drop across that resistor decreases, and the plate voltage rises toward the supply voltage. If part of the plate voltage is returned to the grid through the network, shown in A, the applied grid signal is in phase

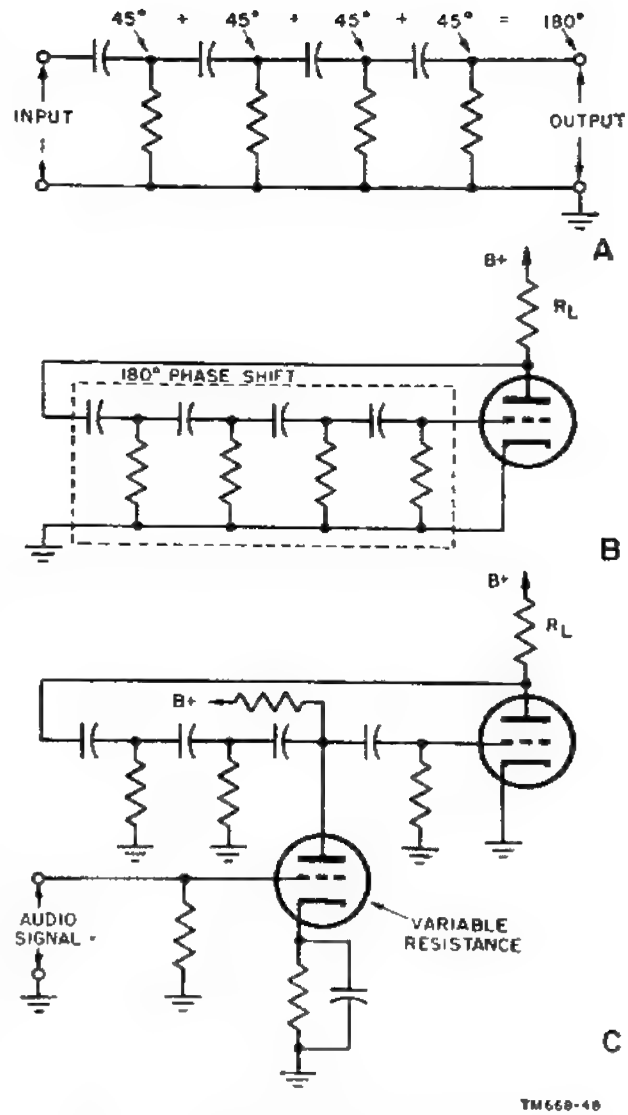


Figure 48. Development of f-m phase-shift oscillator.

with the output of the network. The result is an increase in signal input to the network with correspondingly greater increase in plate voltage, until the accumulative action causes the circuit to break into oscillation. This is the arrangement in the *phase-shift oscillator* in B.

e. The phase-shift oscillator is capable of generating a wide range of frequencies throughout the entire audio range. When the capacitive reactance is such that the phase shift in each section is exactly 45°, the greatest tendency for oscillation exists. Any slight variation in plate current or supply voltage starts the circuit os-

cillating. The frequency of oscillation can be changed by altering the phase shift of any or all of the components of the network. If the resistors are made variable, the frequency can be changed over a considerable range. The amplitude of oscillation is comparatively independent of the losses in the network, provided that the gain of the tube exceeds a certain critical value. Changing the value of one or more of the resistors alters the loss in the network, but does not alter the amplitude appreciably, although it does change the frequency. Therefore, wide frequency excursions can be obtained through changing the resistor values with no variation in amplitude of oscillation. This means that a frequency-modulated signal can be generated which has very little distortion.

f. Instead of varying the resistors mechanically, an electronic circuit has been devised. This is shown in C of figure 48. One of the resistors of the phase-shift network is replaced by the equivalent plate resistance to ground of a vacuum tube. This is accomplished by connecting the plate to the phase-shift network and the cathode to ground. Plate voltage for the tube is supplied through a suitable load resistor. When the grid voltage of the tube is varied by modulating signal, the plate resistance changes over a wide range. In fact, when the tube goes to cut-off, the resistance is extremely high. Similarly, when the grid voltage is zero, the plate resistance is very low. The resistance of this element of the phase-shift network, consequently, can vary over a range which is directly proportional to the modulating signal. This variation of resistance changes the over-all phase shift of the network so that feedback is developed at a different frequency. The variation of plate resistance of the modulator tube is directly proportional to the amplitude of the modulating signal, and the frequency of the phase-shift oscillator is also proportional to the control resistance over a wide range. It follows that the frequency of the oscillator also must be proportional to the amplitude of the modulating signal over an equally wide range. This is the characteristic desired for producing f-m in the subcarrier modulation process.

27. Permeability-Modulated Direct F-M

a. An extremely simple frequency-modulator circuit has been devised for use in very compact portable equipment, where saving an extra tube means an increase in the service life of the batteries. This modulator needs no additional tubes for production of f-m. The operating principle is based on the variation of inductance in an iron core coil with changes in the magnetic flux passing through it. Inductance can be altered by changing the permeability of the iron core—that is, the ease with which the magnetic field passes through it.

b. If the iron core is made a part of an audio transformer in the plate circuit of an audio amplifier, variations in the plate current alter the permeability of the core. This, in turn, changes the inductance of a part of the coil used to form the tank inductance of an r-f oscillator. A fair amount of deviation can be produced in this way without any extra modulating tubes. The system has the disadvantage of high distortion and low deviation capabilities, since the powdered-iron core must be operated in a highly saturated magnetic condition.

Section II. INDIRECT METHODS OF PRODUCING FREQUENCY MODULATION

28. Phase Variation

a. General.

(1) All of the frequency-modulation circuits discussed in section I consisted of oscillators using reactive elements in their frequency-determining networks. Such circuits, however, are inherently unstable in respect to frequency. For very critical applications where it is desired that stations operate on precisely specified frequencies, crystal oscillators are used. A quartz crystal is the electrical equivalent of a resonant circuit with an extremely high *Q*. The resonant frequency is determined by the mechanical properties of the quartz itself.

(2) The reactance changes that can be produced with a direct-method frequency modulator have little effect on the resonant frequency of a crystal oscillator.

Therefore, this oscillator is considered to be fixed in frequency. If a crystal oscillator can be phase-modulated, the output can be converted easily to frequency modulation. Indirect f-m is produced by using a suitable correcting network to convert a phase-modulated signal.

b. Phase Variation. A change in the phase of a signal can be produced by passing the signal through a network containing resistance and reactance. When a voltage is applied to a capacitor and a resistor in series, for example, the current leads the applied voltage by an amount dependent on the relative values of resistance and capacitive reactance. This current develops a voltage across the resistor which leads the applied voltage. If the series combination is considered to be the input, and the output voltage is taken from across the resistor, a definite amount of phase shift is introduced. If the fixed-frequency signal from a crystal oscillator is passed through this network, its phase at the output is shifted by an amount depending on the ratio of the reactance to the resistance. If the resistor can be varied, the phase angle of the network changes to correspond with the newly established ratio of reactance to resistance. When the resistance is varied with an applied audio signal, the phase angle of the output changes in direct proportion to the audio-signal amplitude and produces a phase-modulated signal.

c. P-M to F-M. The basic circuit of a phase modulator with the resistor replaced by the variable plate resistance of a vacuum tube is shown in figure 49. The plate resistance of the triode changes with grid voltage and therefore serves as the variable resistor. Since the plate resistance of the triode varies with the audio signal applied to the grid circuit, the phase between the input of the circuit and the output changes with the audio signal. As the grid swings positive, the plate resistance drops and the phase angle of the output increases; when the grid swings negative, the plate resistance rises and the phase angle decreases. The change of plate resistance with various values of grid voltage is exactly proportional to grid voltage over a small range. If the phase angle of the network is changed between *wide* limits, the *amplitude* of the output changes. This means that the modulator can produce only a limited phase deviation without distortion. In general, it is only reasonably good over a range of less than 25° of phase shift.

29. Constant-Impedance Phase-Shift Modulator

Since the resistance of the modulator tube varies, the voltage across it varies in amplitude

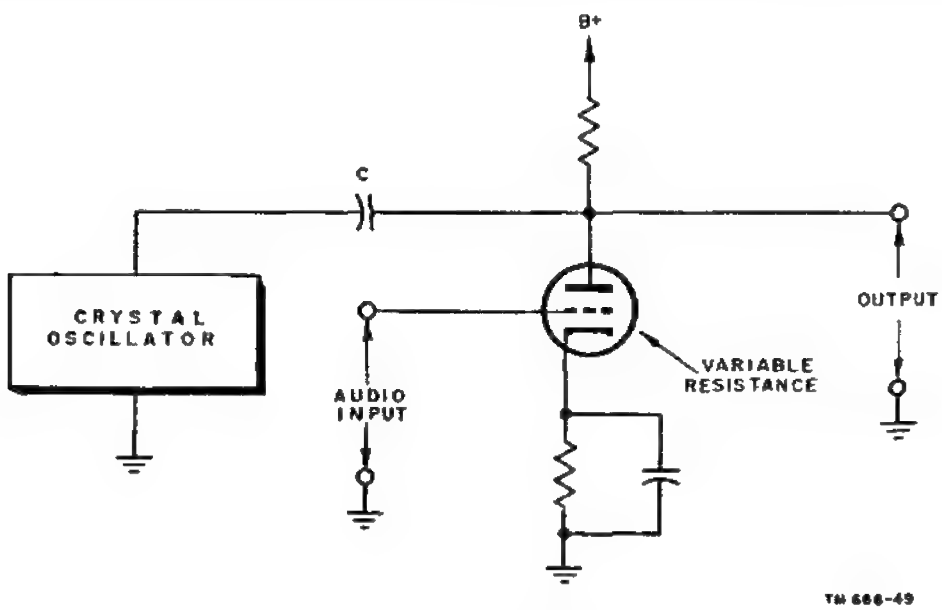


Figure 49. Simple vacuum-tube phase modulator.

as well as in phase. The impedance of the entire phase-shift network varies and introduces distortion in the amplitude of the output signal. This undesirable effect can be overcome, however, if a constant-impedance network is used. A phase modulator having a constant-impedance network is shown in figure 50. The impedance across the cathode resistor, R_k, is used since it changes with variable grid voltage and is more uniform than changes in plate resistance. The resistance between cathode and ground changes with the audio signal, and the phase of the output signal is modulated as desired. The inductor, L, serves to correct any change in the total impedance, keeping the amplitude of the output constant. For any change in frequency, a change in capacitive reactance is canceled by an opposite change in inductive reactance.

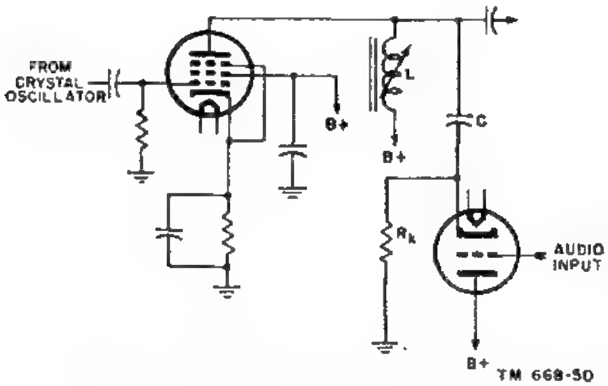


Figure 50. Basic circuit of constant-impedance phase-shift modulator.

30. Audio Correction

a. The equivalent frequency deviation in p-m is proportional to the audio-modulating frequency. This effect is undesirable in indirect f-m production, since the frequency deviation must depend *only* on the amplitude of the audio signal. To convert into f-m the phase-modulated signals produced by the circuits previously described, it is necessary to pass the audio through a correction network to shift its phase 90°. A simple *R-C* series network, with the output voltage taken across the capacitor, can accomplish the desired results. If the resistance is much larger than the reactance of the capacitor at the lowest audio frequency, the current flow is determined primarily by the resistance. Therefore, the current is relatively constant, since the resistance is constant. The reactance of the capacitor changes with frequency, but its effect on the total current flow is small because the reactance decreases as frequency increases and current flow is limited by the high resistance. The voltage across the capacitor, therefore, is equal to the relatively constant current, multiplied by the changing reactance.

b. Since the capacitive reactance is inversely proportional to the frequency and the resistance is fixed, the output voltage is proportional to the reactance alone. Therefore, it also is inversely proportional to frequency, as desired. This type of circuit sometimes is called a *pre-distorter* or *audio-correction* network because it changes the response of the phase modulator so that indirect f-m is produced instead of direct p-m. Since signal amplitude is lost in this type of network, the output sometimes is amplified before it is applied to the phase modulator.

31. Link Phase Modulator

a. The variations in transconductance of a tube with a varying audio signal can be made the basis for a phase modulator. In A of figure 51, the oscillator voltage is applied to the grid circuit of the modulator through coupling capacitor *C*8. The oscillator signal can reach the plate by two paths. One is through the grid-plate capacitance of the tube shown in the diagram as C_{gp}. The other is provided by the transconductance and represents the normal operation of the stage as an amplifier. The plate voltage of an amplifier is always 180° out of phase with the grid voltage. However, the voltage fed to the plate through the grid-plate capacitance is always in phase with the grid voltage. It is therefore 180° out of phase with the amplified voltage.

b. If the tube is operated as a normal high-gain voltage amplifier, the amplified plate voltage is much larger than that caused by the grid-plate capacitance. When a large unbypassed cathode resistor is used, the operating point of the tube can be displaced so that the transconductance and amplification are greatly reduced. The omission of the cathode bypass capacitor permits the variations in plate current to establish degeneration—that is, a varying voltage on the cathode that acts in opposition to effective

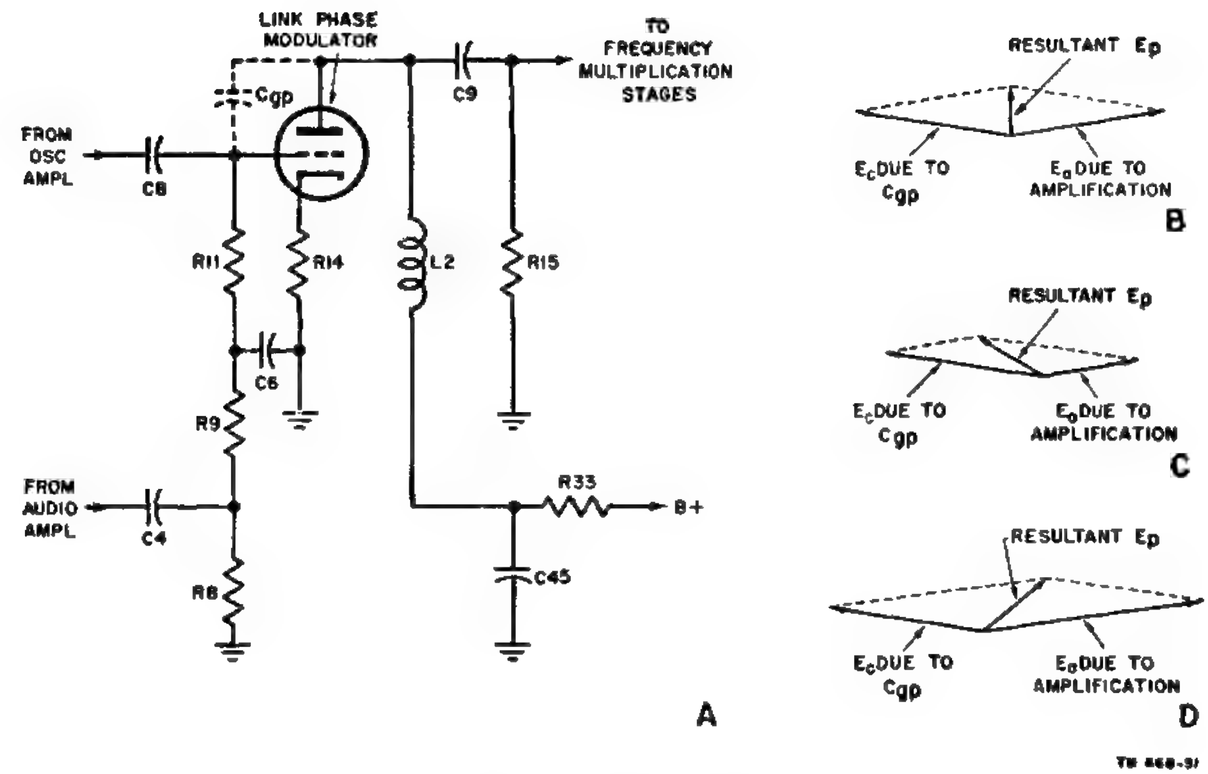


Figure 51. Phase modulator.

grid input voltage—further reducing the transconductance. Because of the grid-plate capacitance, the amplified voltage component on the plate can be reduced in this way to approximately the same magnitude as the component.

c. Resistor *R*8 and the series combination of *R*9 and *R*11 form the grid circuit of the modulator. The plate voltage is applied to the tube through the plate load inductor *L*2, which is broadly resonant with stray capacitances at the operating frequency. The stage is coupled capacitively to the following stage by capacitor *C*9. Capacitor *C*6, in conjunction with the resistance of *R*9, forms the audio correction network.

d. The vector diagrams in B illustrate the relative values and angular relationships of the amplified and capacitive voltage components, E_a and E_c, at the plate. The instantaneous a-c plate voltage is the vector sum of E_a and E_c as shown by the resultant vector, E_p. Audio voltage is applied to the grid of the tube through resistors *R*9 and *R*11, which serve to isolate the r-f and audio circuits. The transconductance is varied at an audio rate by the modulating signal. Consequently, the amplified component of plate voltage varies in amplitude. As the audio signal becomes positive, this plate voltage component is decreased in amplitude but the signal coupled through the grid-plate capacitance from the oscillator does not change in either amplitude or phase. Its vector, E_c, remains constant, as shown in C. Since the amplified voltage, E_a, changes and the capacitive voltage, E_c, does not, the amplitude and phase angle of the resultant, E_p, must change. When the grid of the tube is made negative by the applied audio voltage, the amplified plate voltage increases. The resultant voltage, E_p, changes in amplitude and phase angle, in a direction opposite to its change when the grid swings positive. This is shown in D. The phase of the output signal, therefore, varies in phase, in accordance with the amplitude of the input signal.

e. There are amplitude changes in addition to the phase changes that take place in the r-f voltage on the plate. These are not very great, however, and they can be eliminated in the

stages following the modulator stage. Very little amplitude variation appears in the output of the transmitter.

f. This system is not capable of a phase shift any greater than the approximate 180° separating the amplified and capacitive voltages. In fact, the curvature of tube characteristics at low levels of amplification does not permit the variation in amplified plate voltage to be exactly proportional to the applied audio voltage over its entire range. The actual phase deviation for reasonably low distortion must be much less than 180°. In general, peak phase deviations for the lowest audio frequency to be passed (generally 300 or 350 cps) are about 75°.

32. Sonar Phase Modulator

a. Some of the disadvantages of the Link circuit are overcome in the circuit of figure 52. The oscillator voltage appears on the plate through the dual paths of the tube and the grid-plate capacitance, as before. The resultant instantaneous a-c plate voltage with no modulation is approximately 90° out of phase with the applied r-f grid voltage. When the transconductance is varied by the audio signal, the amplified voltage component varies, and therefore the phase of the resultant also changes.

b. The plate load coil, however, is divided into two parts and the plate-supply voltage is introduced at a tap. The two parts of the coil are 180° out of phase with each other. Part of the out-of-phase voltage from the plate is fed back to the grid through capacitor *C*. The coil is tapped so that the turns ratio between the upper and lower ends is approximately 2 to 1. This provides an adjustment for the phase and amplitude of the voltage acting between grid and plate, by varying the feedback through capacitor *C*. Capacitor *C*1, across the upper part of the coil, varies the magnitude and phase angle of the impedance in the plate circuit through resonance. These circuit modifications appreciably increase the available phase shift by keeping distortion low.

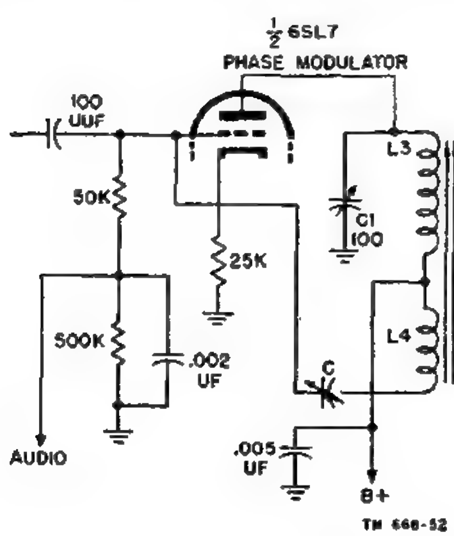


Figure 52. Sonar phase modulator.

c. When the modulating signal voltage is applied to the grid, there is variation in the instantaneous a-c plate current of the tube. Because the coil has a powdered-iron core, the variations in plate current change the magnetic saturation (par. 27). This changes the actual inductance of the coil. With the variation of this inductance caused by the audio signal, the resonant frequency and the phase angle of the plate load circuit also are varied at an audio rate. This phase variation adds to that already produced by the fundamental circuit, and to the variation produced through adjustment of the feedback circuit. The resultant frequency deviation is approximately six times that of the Link phase modulator.

33. Reactance-Tube Phase Modulator

a. It is possible to produce phase modulation by connecting a variable reactance across the resonant load circuit of an r-f amplifier (fig. 53). The variation in the phase angle of a parallel resonant circuit shown in A, is plotted in respect to the frequency. As the frequency of the voltage applied increases, the capacitance in the circuit begins to be predominant, and the resultant total current leads the applied voltage. When the frequency of the applied voltage is lower than the natural resonant frequency of the tuned circuit, the inductance has a lower reactance than the capacitor and, consequently, draws the major share of the current. Therefore, the total current lags the applied voltage.

b. If the frequency of the resonant circuit is varied and the applied voltage kept constant, the same variation of phase angle is produced,

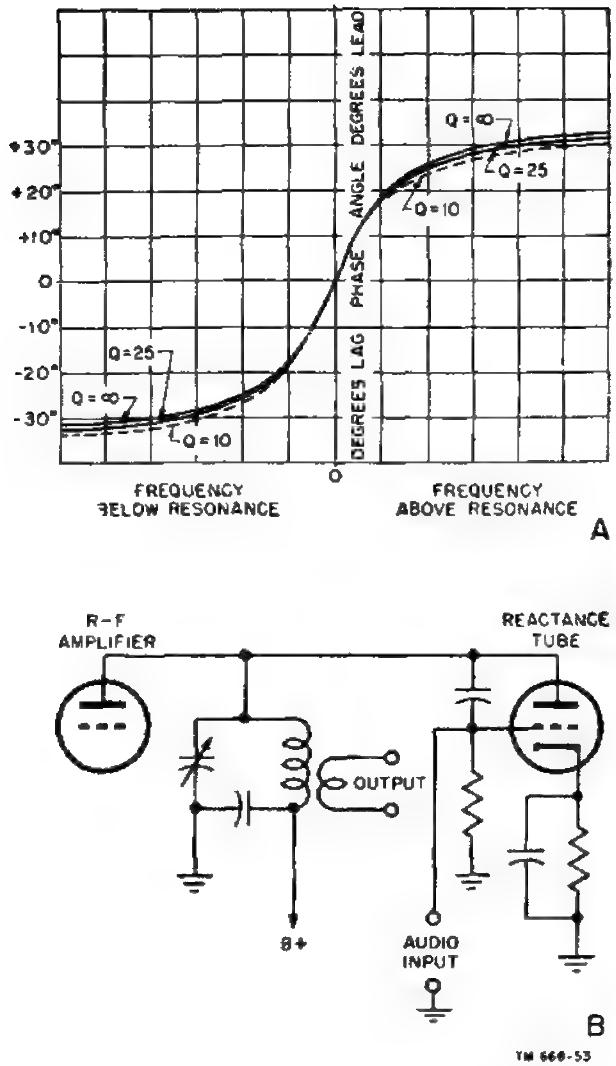


Figure 53. Reactance-tube phase modulation.

The curve in A also applies, except that the center frequency is that of the applied voltage, whereas the frequencies on either side are those to which the resonant circuit is tuned. The circuit can be tuned by changing the value of either the inductance or the capacitance. Changing either at an audio rate with a reactance modulator produces phase modulation. The shape of the phase variation curve depends on the Q of the tuned circuit. This, in turn, depends almost entirely on the construction of the inductor. Therefore, injection of capacitance usually is used to avoid changes in the shape of the curve.

c. The reactance modulator, in B, is designed to inject a variable capacitance across the resonant-circuit load of an r-f amplifier. Note that plate voltage for the modulator and the amplifier are supplied in common. The injected capacitance changes the tuning of the tank circuit in response to the variations in audio signal. This, in turn, changes the phase angle of the current drawn by the tank. Consequently, the phase angle of the output from the tank circuit also varies. The curve in A shows that the change in phase is proportional to the detuning over a small range approximately between the limits of —25° and +25°. This circuit, therefore, cannot produce much phase deviation. It has the advantage of permitting the use of a reactance tube with a crystal oscillator-amplifier combination. However, the wide deviations associated with the reactance modulator when used with a self-excited oscillator cannot be obtained.

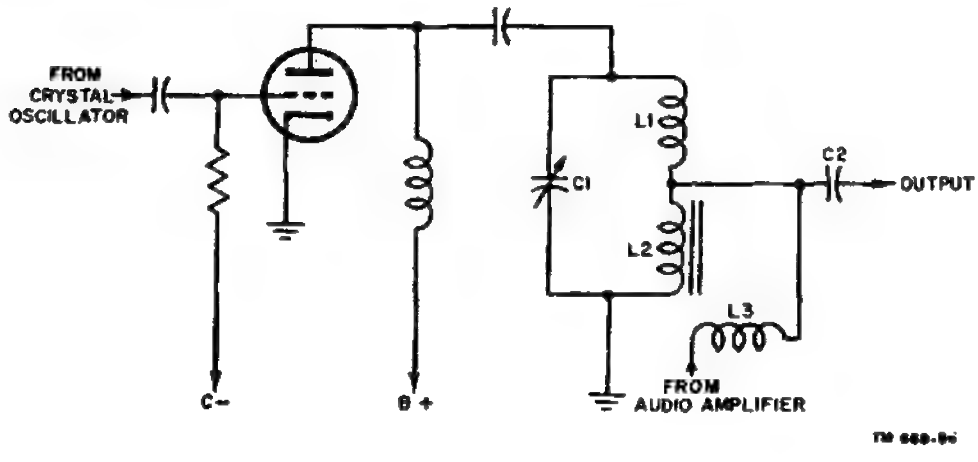


Figure 54. Nonlinear coil modulator.

34. Nonlinear Coil Modulator

a. It is possible to construct a coil which will have the property of introducing phase modulation into a carrier, when both radio and audio frequencies are passed through it. This is termed a *nonlinear coil modulator* (fig. 54). The output from the r-f amplifier is passed through a plate load consisting of the resonant circuit *C*1, *L*1, and the special nonlinear coil, *L*2. The output of the audio amplifier is applied through *L*2 through *L*3, and the phase-modulated signal produced is coupled to the output by *C*2. The rest of the circuit is a conventional r-f amplifier. The nonlinear coil circuit produces a frequency deviation of nearly 1 kc. This phase-modulator circuit is relatively efficient in terms of the amount of initial phase deviation.

b. Normally, when a current is passed through a coil without a magnetic core the current that flows is of the same waveshape as the voltage applied. If, however, a magnetic core is introduced into the coil, this situation changes. When a magnetic field exists in a magnetic material, there is a definite magnetizing force corresponding to that field. As the field increases in strength, the material becomes magnetized until a point is reached where the increase in the magnetizing force produces no increase in the magnetic field set up. When the material is fully magnetized and the magnetic flux cannot increase, a state called *saturation* is reached.

c. When current flows in a coil, it sets up a magnetic field that magnetizes the core material placed in the immediate vicinity. As the current increases, the corresponding magnetic field increases and, also, the magnetization of the core. Because of saturation, however, there is a definite point beyond which any additional current causes no additional magnetization. Special alloys, such as permalloy, when used as cores reach the saturation point at very low values of the magnetizing field. A coil wound around a permalloy core reaches saturation with a very small amount of applied current. When a sine-wave voltage is applied to such a coil, the magnetic flux increases rapidly until the core saturates, and the flux then becomes relatively constant.

d. The relation between the applied magnetizing current and the voltage developed across the coil is shown in figure 55. When the current begins to increase toward its maximum value, the magnetization of the core rises rapidly, with a rapid increase in flux. While the current flow is above saturation (A to B), there is no change in the flux, since the core is saturated. The same situation occurs on the negative half-cycle between points C and D.

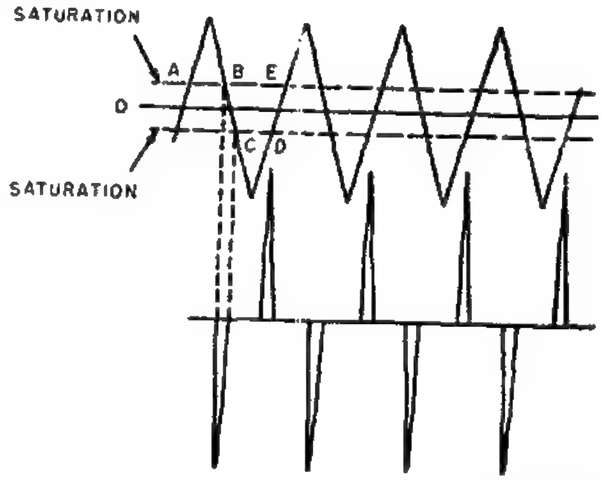


Figure 55. Voltage pulses developed across nonlinear coil.

e. The induced voltage depends on the rate of change of the flux. When the flux is not changing, no voltage is induced. In a coil wound on a permalloy core, the flux is changing rapidly during part of the cycle and, during that time, large voltages are induced in it. These voltage pulses of high amplitudes occur *only* during the periods of rapidly changing flux. At other times, when the flux is nearly constant, little or no voltage is induced across the coil. It is shown in the figure that these voltage changes take place exactly 90° after the current peaks. The polarity of the pulses depends on the direction of the magnetic flux. Therefore, on opposite half-cycles of magnetizing current, the pulses are of opposite polarity. This 90° difference is constant in respect to the magnetizing current, and, since this current is supplied by an r-f oscillator, the pulse is constant in frequency.

f. Assume, however, that, in addition to the r-f energy, audio signals are simultaneously

applied to the nonlinear coil. They have the *same magnetizing effect on the core material* and they combine with the r-f current to produce the current wave, as shown in the first three lines of figure 56. Curve A represents the current in the coil, caused by the carrier r-f. Curve B is the current produced by the modulating signal, assumed to be sinusoidal, for simplicity of analysis. The combined waveform is shown in line C. It is clear that the resultant current no longer goes through the zero axis in the same time interval as before, and, therefore, the region of maximum rate of change of flux is different for each cycle and depends on the audio voltage. These combined currents produce voltage pulses across the coil at different instants during the audio cycle, as in D. The variation in the level of the r-f current at different points of the r-f cycle causes this effect. Sometimes, the pulse is produced at the normal interval of the unmodulated carrier. At other times, the pulses are spaced more, or less, than 360° part. These variations in the spacing of the pulses, with different values of modulation voltages, are *obviously equivalent to displacements in the relative phase of the pulses*. In other words, a change in the a-f voltage shifts the phases of the pulses in respect to the phase of those pulses produced by the unmodulated carrier. Therefore, these pulses are effectively phase-modulated.

g. If the phase-modulated pulses of voltage derived from the nonlinear coil are applied to a rectifier and limiting amplifier, only the pulses of one polarity will be passed. Furthermore, the *limiting action will reduce the slight variations* in the amplitude of the pulses that appear in D. The action of this rectifier and limiter is shown in E. When pulses of sharp amplitude pass through a resonant circuit, they set it into oscillation at its natural resonant frequency. If these pulses from the output of the rectifier pass through a resonant circuit, which is tuned to their repetition frequency, a sine wave is produced. Since the pulses are phase-modulated by the audio voltage, the resultant sine wave also will be phase-modulated.

35. Balanced Modulator

a. Vector Relations Between Amplitude and Phase Modulation—Amplitude Modulation. In the earlier chapters, it was pointed out that *when the modulation index of an f-m signal is* less than .5, only one effective side-band pair is produced. Therefore, the side bands of an f-m wave with a low modulation index are similar to those of an a-m wave. The major difference between the side bands of a normal a-m wave and those of an f-m signal with a small deviation lies in their phase relationship to the carrier. An a-m signal consists of a carrier and two side bands for each audio frequency present in the modulating wave. For an f-m wave with modulation index of less than .5 the same holds true. If the two side bands of the a-m signal are added together vectorially, a new wave results, called a *double side-band wave*. It represents the difference between the frequency of the modulated wave and that of the carrier. For a narrow-band f-m signal, a similar double side-band wave also is formed when the two side bands are added. The a-m and f-m double side-band waves are ordinarily identical with one another, if they are of the same frequency and amplitude.

b. Double Side-Band Vectors. In the vector diagram shown in A of figure 57, the vector, *AB*, represents the amplitude of the unmodulated carrier. Vectors *BC* and *BE* are the two side-band components. *BD* is the vector sum of the side bands, or the double side-band vector. The resultant a-m wave therefore is the vector, *AD*. Each of the side-band vectors rotates about the tip of the carrier vector with a rotational frequency equal to the difference in frequency between the carrier and each side band, respectively. Consequently, the two side-band waves, which are equally spaced in frequency on either side of the carrier, rotate at the same speed but in opposite directions. One is higher in frequency, producing a positive difference, whereas the other is lower than the carrier frequency, producing a negative difference. The resultant of the two side bands, therefore, must move on the same line as the vector representing the carrier frequency, if it starts in phase with the carrier. This either adds to or subtracts from the amplitude of the carrier, as shown in B and C. In B, the two vectors add to produce a resultant that adds to the carrier amplitude. In C, they add in the opposite direction, and therefore subtract from the carrier vector. This

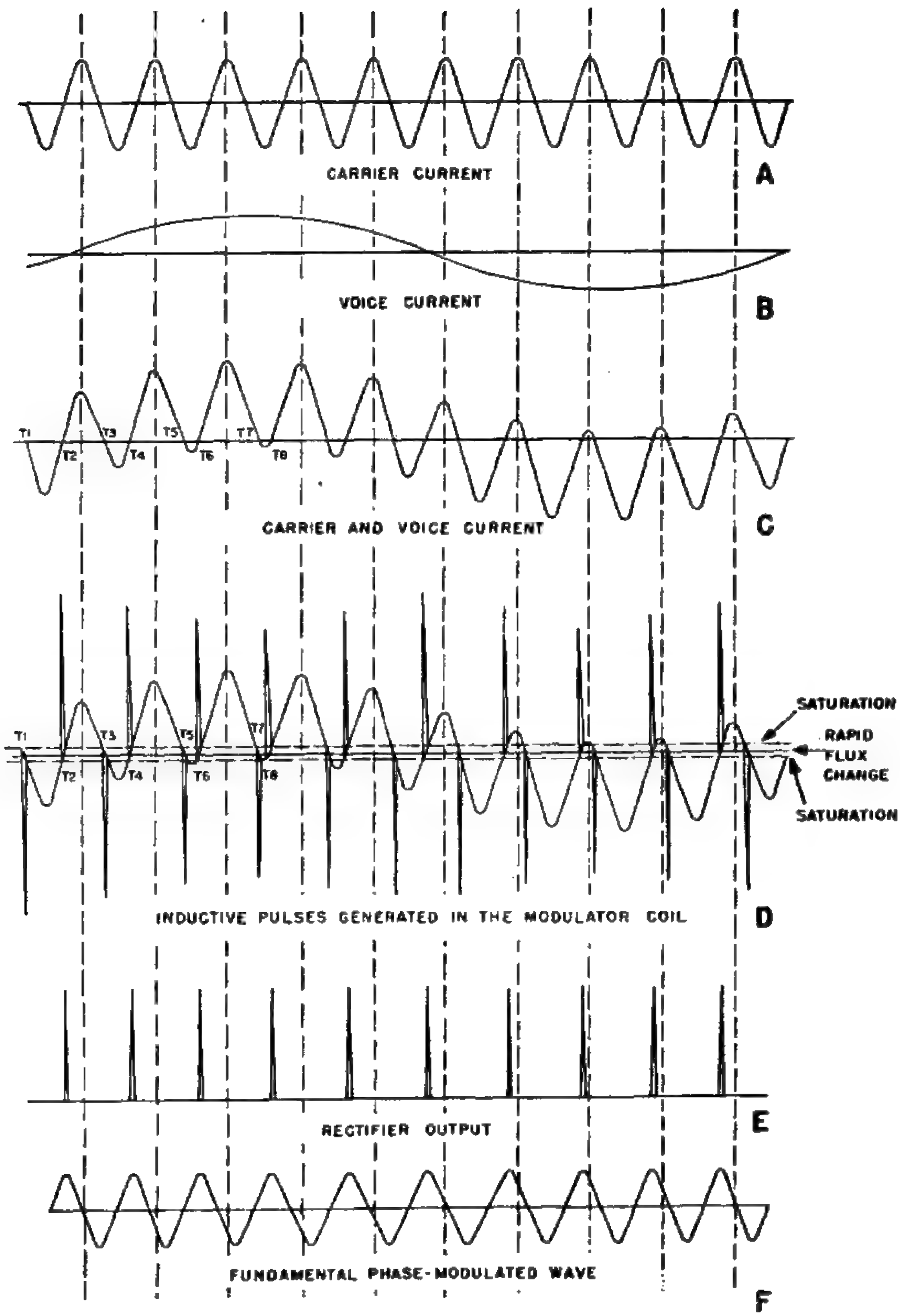


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Figure 56. Waveforms developed in nonlinear-coil modulator.

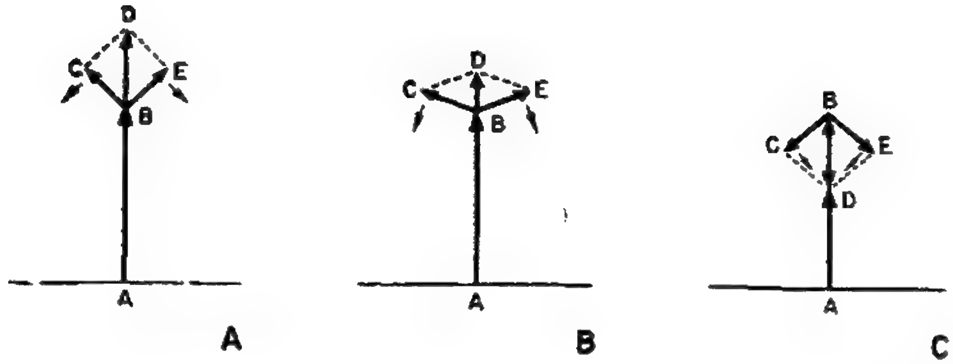

Figure 57. Vector relations in double side-band a-m system.

causes changes in the amplitude of the combined signal, but there are no phase changes in the carrier itself. The resultant of the combined side-band signals is in phase with the carrier frequency.

c. Vector Relations Between Amplitude and Phase Modulation—Phase Modulation. In the case of phase modulation, when the frequencies and amplitudes of the side bands are the same as those in a-m (modulation index less than .5), the vector relations of the side bands and the carrier are different. Specifically, the double side band, which results from the addition of the two rotating side-band vectors, is such that it is always 90° out of phase with the carrier, as in A of figure 58. Instead of starting in phase with the carrier, as they do in a-m, the side-band vectors begin to rotate in opposite directions, but 90° out of phase with it. Consequently, the double side-band vector is always 90° out of phase with the carrier. It can be either on the right side of the carrier, as in B, or on the left side, as in C. The resultant of the sum of the two side bands and the carrier (vector *AD*) therefore shifts alternately from one side of the unmodulated carrier position to the other. It is undergoing a variation in respect to the unmodulated phase condition of the carrier, and, consequently, p-m is produced. When the resultant, *AD*, is on the right of the carrier, an angle, $\phi 1$, is generated. As the side bands rotate, and *BD* becomes smaller, vector *AD* changes, generating angle $\phi 2$. Similarly, $\phi 3$ is the generated angle, when the double side-band resultant is on the lower side of the carrier.

d. Changing A-M to P-M. When the deviation is low, a-m, f-m and p-m signals are all similar, with the exception of the phase relationship of the double side band to the carrier. This indicates a possible way to generate an indirect f-m signal from an a-m wave. A signal is amplitude-modulated with a low percentage

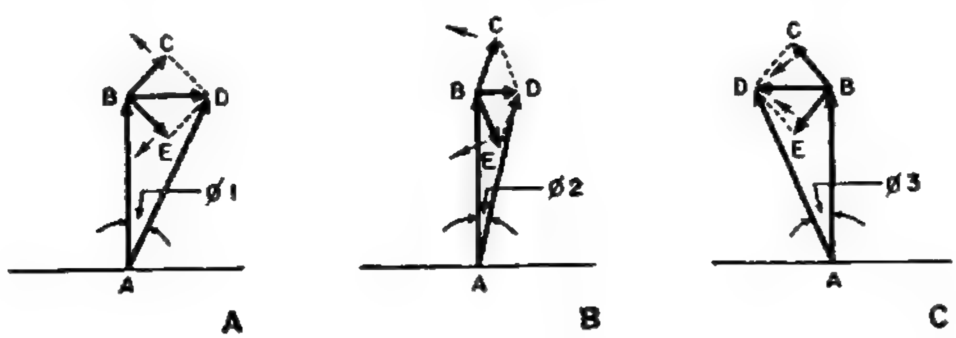

Figure 58. Vector relations in a double side-band f-m system.

of modulation. Then it is passed through a circuit which removes the carrier-frequency component, leaving only the double side band. This double side band is shifted in phase by 90°. If it now is recombined with the carrier component, the result is a phase-modulated wave. This amounts to converting the vector diagram in A of figure 57 into that of A of figure 58 by rotating the double side band 90°. When this is done, the resultant vector varies in phase and in amplitude simultaneously, in respect to the carrier frequency. The slight amplitude variation remaining can be removed in a limiting amplifier, as explained previously.

e. Indirect F-M System: A-M to P-M.

(1) A system for the production of f-m by the indirect method is shown in figure 59. A crystal-controlled oscillator generates a low-frequency carrier that is fed into an r-f amplifier. At the same time, it passes through a device called a balanced modulator which amplitude-modulates it with an audio wave that has been passed through an audio correction network, producing the two side bands and simultaneously removing the carrier itself. The output from the balanced modulator then is fed through a network that shifts the phase of the double side band by 90°.

(2) The amount of amplitude modulation developed by the balanced modulator is small and the resultant side bands are of low amplitude. The double side band, after passing through the 90° phase-shift network, is recombined with the carrier frequency as derived directly from the crystal oscillator. This results in a signal that is almost pure p-m, because the original amplitude variations of the double side bands are small in comparison with the unmodulated carrier. The amount of phase deviation possible with this system is small because of the low amplitude of the double side bands. However, when phase deviation is kept small, distortion in the modulation is extremely low.

f. Balanced Modulator.

(1) The heart of the f-m system outlined above is the circuit that modulates the carrier wave, producing the double side band while eliminating the carrier itself. There are several forms of such balanced modulator circuits,

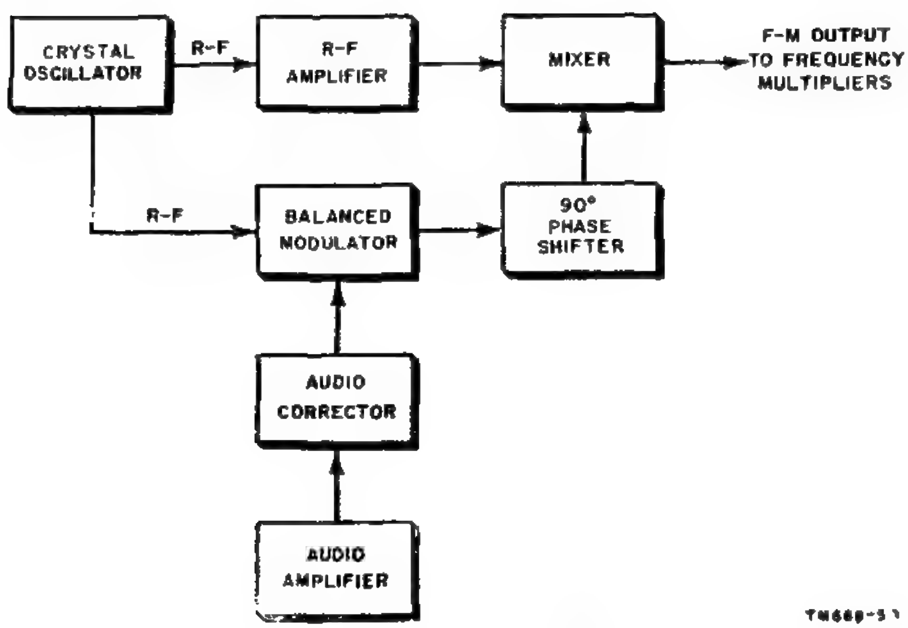

Figure 59. A-m to p-m to f m—indirect method.

one of which is shown in figure 60. In this circuit the output from the crystal oscillator is coupled to the control grids of two tetrodes, from either side of a tuned transformer. Because of this method of coupling, the voltages at the grids of the tubes are 180° out of phase. The plates of these tubes are connected in parallel and an output is developed across a common load impedance, usually a parallel-resonant circuit.

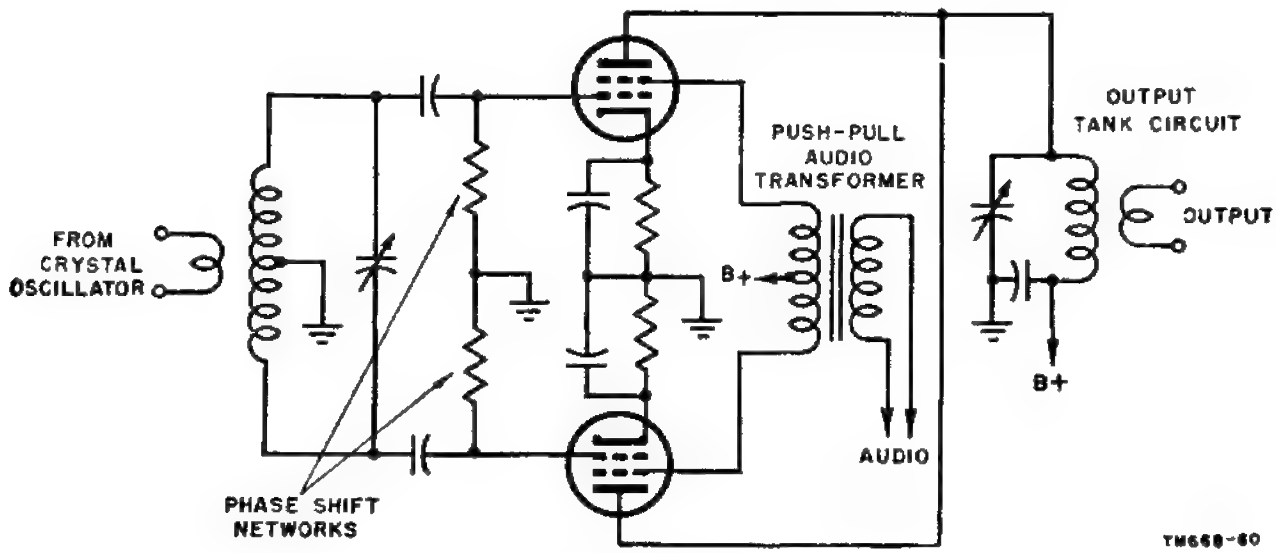

Figure 60. Balanced modulator.

(2) The grid excitation voltage passes through a resistor-capacitor network in which the reactance of the capacitor is much larger than the resistance. Therefore, the voltages at the grids, in addition to being out of phase with each other, are 90° out of phase with the voltage produced by the oscillator. If the two modulator tubes are identical, the variation of one grid voltage in a positive direction and the other in the negative direction produces equal and opposite changes in voltages at the two plates. Because the plates are connected in parallel, and the equal and opposite changes cancel, the input carrier is effectively eliminated across the common plate load. The voltage from the audio circuits is applied to the screens of the two modulator tubes causing a variation in plate current which is proportional to the modulating voltage on each screen grid. The audio signal unbalances the modulator tubes and therefore the side bands produced by the audio and the carrier appear in the output. The situation can be understood in terms of the audio voltages applied to the screen grids. Since the input is positive on one control grid and negative on the other, at any instant of time, and then reversed on the following half cycle, first one tube conducts and then the other. Consequently, output always will appear in the plate circuit from the tube which is conducting, whereas the other is almost at cut-off. When the screen of one tube is made more positive than the other by the audio voltage, the output increases. This action produces side bands equal to the sum and difference of the audio signal and the carrier, whereas the carrier is effectively canceled.

g. Other Balanced Modulator Circuits—Parallel-Grid Push-Pull Plate Circuit. It also is possible to produce the double side band without the carrier by connecting the control grids of the two tetrodes in parallel and the plates

in push-pull, applying the audio voltage to the screens, as before. With in-phase voltages on the control grids, equal voltages are produced at each plate. If these voltages are combined in a transformer, they are in phase at its opposite terminals. A transformer produces no output unless the voltages are out of phase on each side; therefore, the carrier cannot pass through the circuit. However, the voltages applied to the screens are in push-pull; that is, they are out of phase with each other. When the audio voltage is applied to the screen grids, the circuit is unbalanced, as before, and current flows in the output circuit. The 90° phase shift that is needed is incorporated in the output circuit by using a suitable network of coils and capacitors. It also is possible, by using the angle of lag in the current that flows in an untuned secondary, to make the output transformer itself act as a phase shifter.

h. Ring Modulator. A simple balanced modulator (fig. 61) that requires no vacuum tube can be constructed from four *varistors*. Varistors are special resistors made of material whose resistance *varies* with the applied voltage. In general, the resistance decreases as the voltage increases, so that the current flow increases at a rate much faster than in an ordinary resistor. The action is much like that of a simple vacuum-tube diode. Four copper-oxide rectifiers connected in a bridge circuit and sealed in a metal container commonly are called a varistor. The carrier is applied at two terminals, and the audio and load are applied to the other two. The carrier is balanced out in the bridge, and the unbalanced audio current that flows in the varistors produces the side bands in the output. These devices are used most

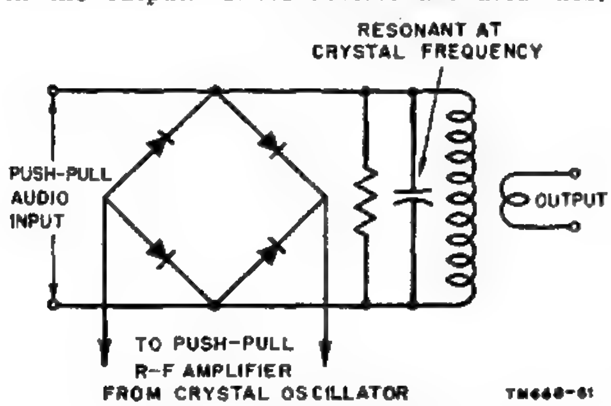

Figure 61. Varistor-ring modulator.

widely in telephone work for the production of a double side-band pair, where they have the advantage of not requiring any currents besides those of the carrier and the modulating signal for their operation. If the double side band is recombined with the carrier, after a 90° phase shift, p-m is produced.

36. Modified Balanced Modulator

a. The balanced modulator and 90° phase-shift arrangement is satisfactory in terms of distortion. However, the deviation produced is very low, and a modification of the basic circuit has been worked out which does not require balancing out the carrier completely, shifting the side bands 90°, and reinserting the carrier. This circuit is used in many types of mobile and fixed equipment where simplicity and reliability are necessary. Basically, it is a cross between the Link modulator and the balanced modulator. A simplified version of this oscillator-modulator circuit is shown in A of figure 62. The grids of the pentagrid modulator tubes are excited through an R-C network in the output circuit of the oscillator. This network starts with a parallel-tuned circuit resonant at the crystal frequency. The opposite sides of the coil are 180° out of phase. This out-of-phase voltage is passed through a very small capacitor, *C*106, and the output is taken across a small resistor, *R*104. This introduces an additional 90° phase shift between one side of the coil and the other, for a total of 270°. The result is that the opposite grids of the modulator tubes are 270° out of phase. They are connected across a coupling network consisting of two capacitors, *C*104 and *C*107, and two resistors, *R*105 and *R*106. *C*105 serves to ground the center point between the two resistors. The modulator grids therefore are 45° out of phase in respect to the reference phase and 270° out of phase with each other.

b. In the vector diagram shown in B, the two 90° out-of-phase input signals, E_1 and E_2, are combined in the output of the modulator tubes to produce a resultant, designated as E_R. The two number 4 grids are fed by a push-pull audio signal. When the audio voltage on the grid of one tube goes negative, it reduces the transconductance of that tube. This, in turn, reduces r-f voltage and, correspondingly, the length of the

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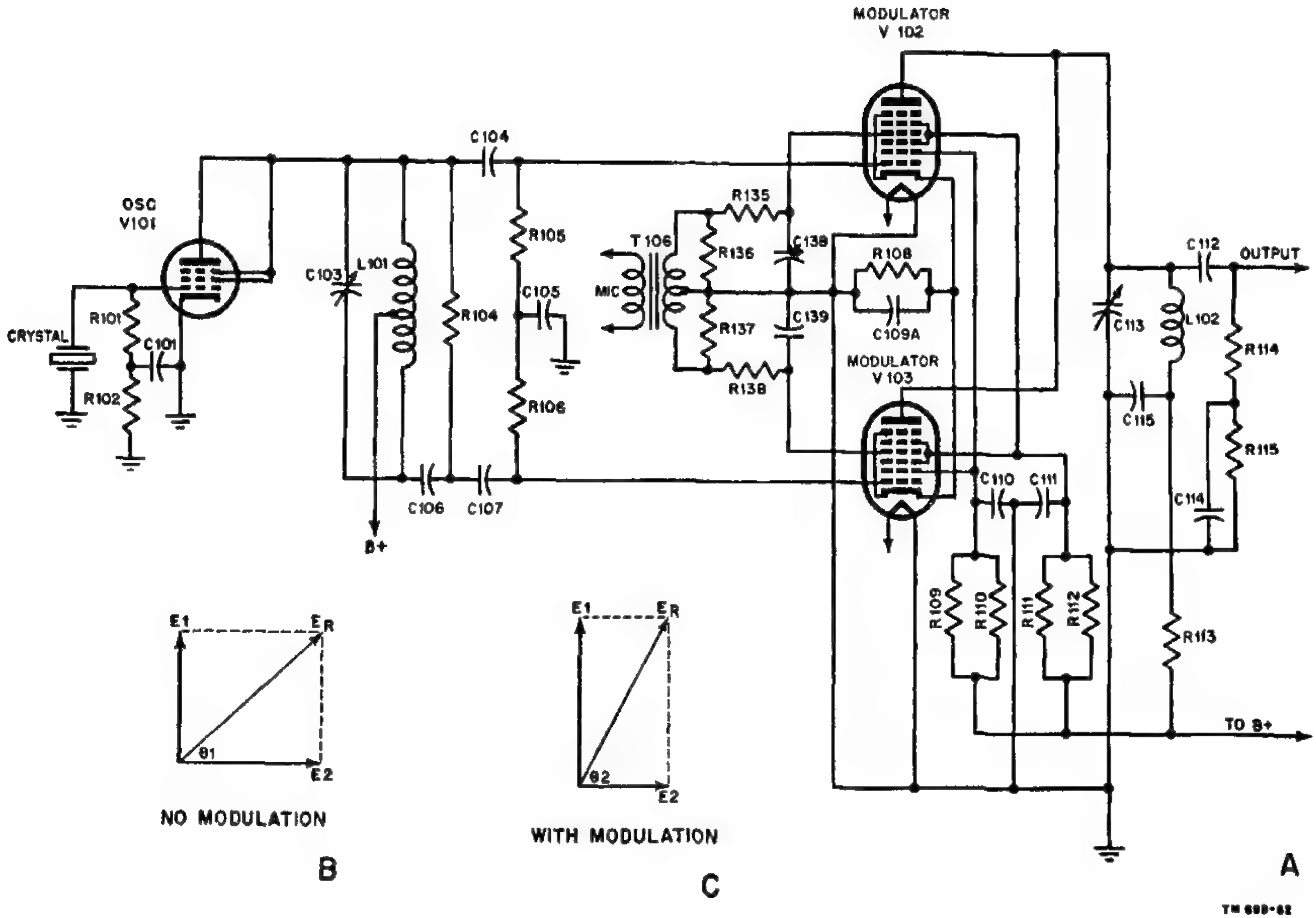


Figure 62. Modified balanced modulator.

vector representing the r-f voltage. Simultaneously, the other tube is being driven to a higher r-f output voltage by the positive part of the audio cycle applied to its grid. Therefore, the vector representing the increase in output from the second tube lengthens, as shown in C. The resultant of the shorter vector of plate voltage of one tube and the longer vector of the other swings from side to side, and a considerable amount of phase deviation is introduced. This deviation can never exceed the 90° difference in the r-f input signal to the two tubes. However, the distortion that is encountered usually dictates a lower value of phase deviation than indicated. The circuit resembles the balanced modulator very closely at first glance, but the difference between them lies in the fact that in this arrangement the grids are fed *90°* out of phase, and there is *no* cancellation of the carrier in the parallel plate circuit. In the balanced modulator, the grids are *180°* out of phase and the carrier *is* balanced out. Inherently, this circuit is not capable of the low distortion of the true balanced modulator carrier reinsertion arrangement.

Section III. SUMMARY AND REVIEW QUESTIONS

37. Summary

a. F-m always is produced near the frequency-determining section of the transmitter at a low power level.

b. Varying the inductance or capacitance in an oscillator in accordance with an audio signal produces f-m.

c. A reactance tube injects capacitance or inductive reactance into an oscillator circuit.

d. The transconductance of a vacuum tube is the ratio of a small change of plate current to a small change of grid voltage, with plate voltage held constant.

e. The amplification factor is the ratio of a small change in plate voltage to a small change in grid voltage, with the plate current held constant.

f. The plate resistance is the ratio of a small change in plate voltage to a small change in plate current, with the grid voltage held constant.

g. The product of the plate resistance and the transconductance equals the amplification factor.

h. The basic expression for the resistance and reactance injected by a reactance modulator is

$$Z = \frac{1}{g_m} + \frac{1}{g_m} \times \frac{Z_a}{Z_b}$$

where Z_a and Z_b are the impedances of the voltage divider connected between plate and grid, and between grid and cathode, respectively.

i. A reactance tube can inject capacitance or inductance, depending on the components of the voltage divider.

j. The input impedance of a tube changes with plate load or transconductance, and this change is known as the Miller effect.

k. A diode modulator acts as a variable resistor in series with a capacitor placed across a tank circuit, effectively changing the phase angle of the current drawn from it, and thus changing the frequency of the oscillator.

l. An R-C oscillator can be frequency-modulated by varying one of the frequency-determining resistors. Variation is accomplished by replacing a resistor with the dynamic plate resistance of a tube.

m. Crystal oscillators are more stable than conventional vacuum-tube oscillators, but they cannot be frequency-modulated directly.

n. A simple phase modulator introduces a variable time delay in a circuit, through which the carrier signal must pass. The variable time delay can be produced across one element of an R-C series circuit. The resistor can be replaced by the plate resistance of a tube.

o. A-m as well as p-m is produced in the simple phase modulator; therefore, the constant-impedance characteristic of a tuned circuit is used to overcome this disadvantage.

p. An R-C series circuit, with output taken across the capacitor, is used as an audio corrector to convert p-m to f-m.

q. The two ways by which a signal can reach the plate of a tube are grid-plate capacitance and tube transconductance.

r. The signal path, via the transconductance, can be varied by grid signal changes. This produces a variable voltage on the plate that combines with the out-of-phase capacitive voltage to produce a variable phase shift.

s. Greater phase shift can be obtained by feeding back part of the output into the grid circuit. An iron-cored coil in a tuned plate circuit also increases the effective deviation.

t. It is possible to produce phase modulation by varying the phase of a parallel resonant circuit by detuning it with the injected reactance of a reactance modulator. This parallel resonant circuit is used as the plate load in an r-f amplifier.

u. A nonlinear coil can be used to generate a series of phase-modulated pulses when both r-f and a-f currents flow through it at the same time.

v. When the modulation index is low, there is only a very slight difference between an f-m and an a-m signal.

w. The vector sum of the two side bands of a narrow-band f-m signal, or ordinary a-m signal, is known as the double side band. In f-m, the double side band is 90° out of phase with the carrier, whereas in a-m it is in phase with it.

x. A-m can be changed to p-m by removing the carrier, rotating the double side band 90°, and reinserting the carrier.

y. The double side band can be produced and the carrier rejected by a balanced modulator.

z. The grids of a balanced modulator are operated either in push-pull or in parallel, with the plates connected oppositely. The audio signal unbalances the modulator, permitting the side bands to go through.

aa. Varistors, arranged in a bridge circuit, can be used to produce a double side band without the carrier.

ab. In the modified balanced modulator, the grids are excited 90° out of phase. Considerably more deviation is obtained in comparison with the ordinary balanced modulator.

38. Review Questions

a. At what power level is f-m produced?

b. What varies the frequency of an oscillator?

c. What is the purpose of a reactance tube?

d. What are the three basic characteristics of a vacuum tube?

e. What is the relationship between them?

f. What are the four combinations possible for the voltage divider in the reactance modulator?

g. What is the Miller effect? Describe how it can be used to frequency-modulate an oscillator.

h. Describe the operation of a diode modulator.

i. How can an R-C oscillator be frequency-modulated?

j. What is the disadvantage of a conventional oscillator as compared with a crystal oscillator in an f-m transmitter?

k. What purpose does audio correction serve?

l. What happens to the phase of a parallel-resonant circuit, operating above its resonant frequency? Why?

m. Why must the amplitude of the r-f carrier, alone, be sufficient to produce saturation in the nonlinear coil modulator?

n. How are the phase-modulated pulses converted into continuous waves?

o. When is an f-m signal nearly equivalent to a-m?

p. What is the double side band?

q. What is the phase of the double side band, relative to the carrier, in f-m? In a-m?

r. How is an a-m signal converted into an f-m signal?

s. How are the input and output circuits of a balanced modulator connected?

t. How do the side bands pass through the balanced modulator?

u. Why is a 90° phase shift network put in the grid circuit of some balanced modulators?

CHAPTER 4

F-M TRANSMITTER

Section I. BASIC CIRCUITS

39. Basic Transmitter

a. The various direct and indirect methods for producing frequency modulation involve changing either the frequency or phase of an oscillator in accordance with some modulating signal. In the direct method, the modulating signal is injected into a modulator whose output varies the frequency of the oscillator in accordance with the original modulating signal. In the indirect method, the modulating signal is passed through a correction network to a phase modulator. The correction network changes the phase of the modulation in such a manner that, when the output of a crystal oscillator is passed through the modulator, the oscillations are frequency-modulated in accordance with the modulating signal.

b. It is extremely important that the transmitter be transmitting at its designated frequency. To achieve maximum frequency stability, therefore, the oscillator is operated at relatively low frequencies, and the lower the frequency, the more stable the oscillator. This means that the center frequency of the f-m signal output of the modulator-oscillator section is lower than the carrier frequency desired for transmission. To raise the f-m signal to the correct frequency, it is passed through a series of *frequency multipliers.* Each stage of frequency multiplication raises the frequency of the signal input by some multiple of the fundamental frequency: A doubler produces a signal at its output that is twice the frequency of the input signal; a tripler raises the frequency three times; a quadrupler four times. When the input to the frequency multiplier is an f-m signal, the multiplier produces an increase in the frequency deviation.

c. The oscillator-modulator and frequency-multiplier sections of the transmitter are operated at low power levels, and the output of the final multiplier is too weak to be transmitted. A power amplifier similar to those in a-m transmitters acts as the final stage in the f-m transmitter to build up the signal to the power level desired.

d. The indirect method of f-m transmission generally uses a crystal oscillator to produce the r-f signal to be modulated since the crystal increases the frequency stability. In direct f-m, a crystal oscillator with its fixed frequency cannot be used, because the oscillator must be free to change frequency in accordance with the modulating signal. Whatever the type of direct f-m oscillator, it cannot be made as stable as a crystal oscillator, and an auxiliary circuit must keep it on the correct center frequency. Such a circuit is called *afc,* or *automatic frequency control.* A complete description of the f-m transmitter covers all the circuits mentioned above. Regardless of the circuits used in a particular transmitter, however, it must be remembered always that the *sole* purpose of the equipment is the transmission of intelligence from one point to another.

40. Frequency Multiplication

a. General. In f-m transmitters, frequency multiplication of the f-m signal performs two functions: It increases the frequency of the signal to the value desired for transmission, in this way acting the same as a frequency multiplier in an a-m transmitter. It also increases the effective frequency deviation of the f-m signal.

b. *Increasing Frequency Deviation.*

(1) The f-m signal from the oscillator-modulator section has a center frequency f_c and a frequency deviation of Δf caused by the modulating signal. The f-m signal therefore varies from a maximum of $f_c + \Delta f$ to a minimum of $f_c - \Delta f$. For example, with an oscillator whose unmodulated output frequency is 100 kilocycles, a certain audio signal causes this frequency to swing between 95 and 105 kc, and the frequency deviation therefore is ± 5 kc.

(2) If this f-m signal is impressed on the grid of a tube which is operating as a doubler, the center frequency of the doubler output is twice f_c, or 200 kc. Since the multiplier doubles any frequency appearing at the grid, when the f-m signal is deviated to 95 kc, the output frequency is 190 kc. When the f-m signal is 105 kc, the output is 210 kc. The multiplier output therefore varies from 190 to 210 kc, and the new deviation is 10 kc. By doubling the frequency at its input, the multiplier also has doubled the frequency deviation. The amount of multiplication used depends on the frequency to which the signal must be raised and the amount of frequency deviation desired. The greater the deviation, the greater is the bandwidth of the f-m signal transmitted.

c. *Basic Frequency Multiplier.*

(1) Essentially, frequency multipliers are harmonic generators; that is, the output frequency is some multiple of the input frequency. The output circuit must contain not only the original input frequency, but also harmonics of it, and is made selective to the harmonic desired, all other frequencies being rejected. To produce these harmonics, a class-C amplifier stage is used with its plate circuit tuned to the frequency of the harmonic desired. Such an amplifier is called a frequency multiplier.

(2) Figure 63 is the simplified schematic of a frequency doubler using a tube operated in class C. The input signal is impressed across the grid circuit through transformer *T1*. The frequency of this signal is the fundamental frequency, *f*, of the system. Because of class C operation, the plate current is nonlinear and therefore rich in harmonics. If a tuned circuit, *L2-C2*, is placed in the plate circuit, it can be tuned to the frequency of the desired harmonic. When *L2-C2* is tuned to twice the frequency of *L1-C1*, this stage becomes a doubler.

(3) In figure 64, a typical i_b-e_c characteristic curve for a tube in class C operation is shown. The grid is biased far below cut-off, and plate current flows only during the portion of the signal that takes the grid voltage above cut-off. This portion of the signal is only a fraction of the positive half-cycle, and the resultant plate-current flow is in the form of short pulses during the

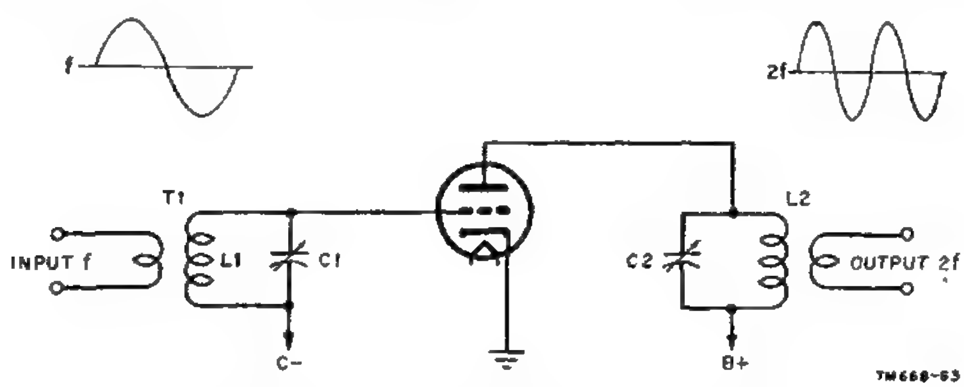


Figure 63. Simplified schematic of frequency doubler.

time the grid voltage is above cut-off. The amplitude of these pulses depends on the amplitude of the signal at the grid.

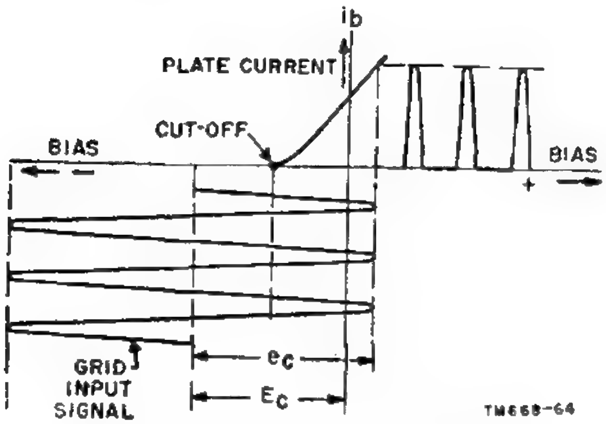


Figure 64. Class C amplifier operation.

(4) The output pulses of plate current are distinctly nonlinear in respect to the input signal, and it is this characteristic of class C operation that is used to produce an output at the harmonic frequency. If the output waveform exactly reproduced the sinusoidal input waveform, then no frequencies could be present in the output that were not present in the input. If the *shape* of the waveform is changed, new frequencies must have been added. These new frequencies take the form of harmonics of the input frequency. The frequencies present depend on the duration of the pulse and how sharply it is peaked. The shorter and sharper the pulse, the more harmonics are produced in the output. With a doubler, for example, the tube is biased farther beyond cut-off than in a class C amplifier. This produces a shorter and sharper pulse which contains sufficient energy at the second harmonic to drive the tank circuit. Biasing the tube farther beyond cut-off requires a larger input signal to produce the same amplitude of current flow in the plate circuit.

(5) The amount of multiplication depends on *the final frequency and frequency deviation desired.* A single multiplier can be used to multiply the frequency by as much as five times. The higher the order of multiplication, however, the lower the output of the stage. Usually, the desired multiplication is obtained through several successive stages, the highest-order multiplier used being the quadrupler. For example, the center frequency of the signal from a modulator-oscillator section is 10 megacycles, and it is desired to transmit this signal at a center frequency of 80 mc. This means multiplying the signal frequency eight times. Two possible ways of obtaining this amount of multiplication are shown in figure 65. In A, three multiplication stages are used. Each stage is a doubler with its output tank tuned to the second harmonic of the signal input at its grid. The first stage raises the center frequency of the signal from 10 to 20 mc. This is applied to the center frequency of the input of the second doubler, which raises the center frequency to 40 mc. The last doubler stage raises the signal to the desired center frequency of 80 mc.

(6) In B, the same result is obtained by using a frequency quadrupler followed by a doubler. The frequency is multiplied four times in the first stage, and the output then is fed to a doubler. This produces the same amount of multiplication as the three doublers used in the first method, and with only two tubes. The second method is used where compactness and economy are desirable, but efficiency and power output are lower.

d. Push-Push Doublers. When the plate tank of the circuit in figure 66 is tuned to twice the frequency of the input, the circuit acts as a very efficient doubler. Its operation can be considered similar in action to that of a full-wave rectifier. The grid of each tube is biased approximately to cut-off so that plate current flows in each tube on succeeding half-cycles. When the signal input across the secondary of *T* makes the grid of *V*1 positive in respect to its cathode, the tube conducts; at the same time,

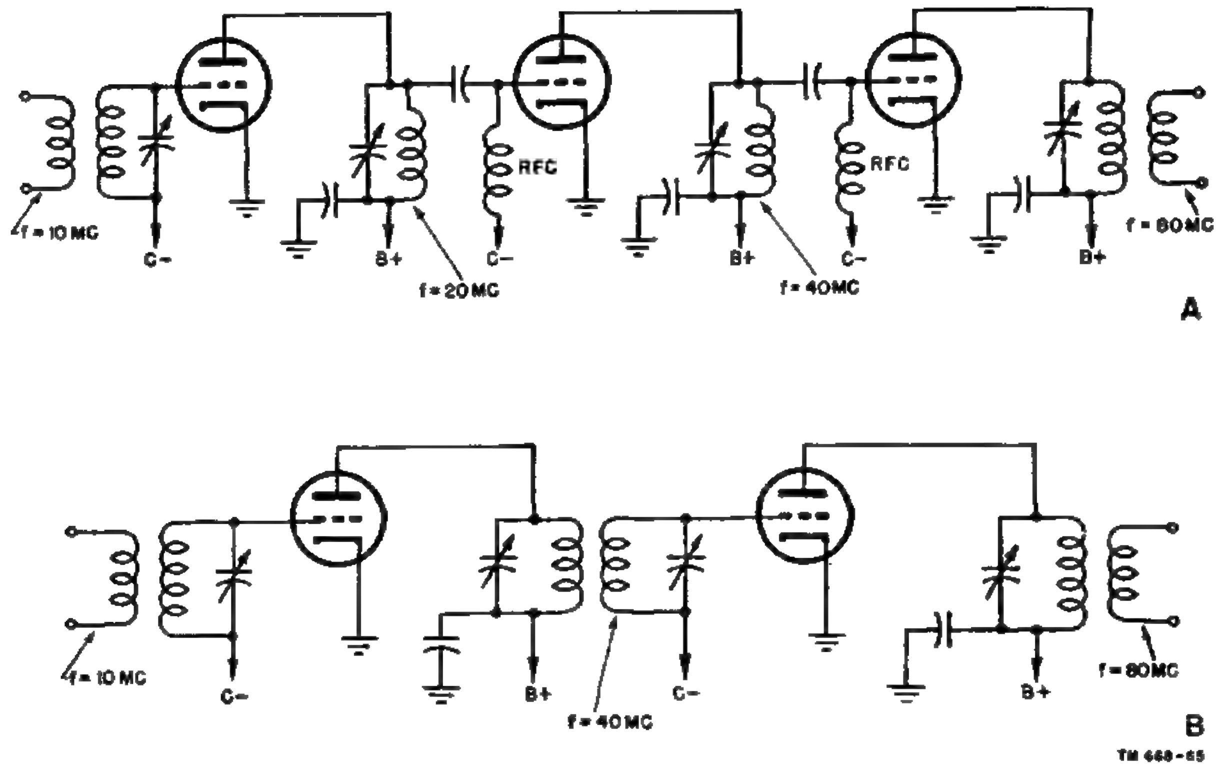


Figure 65. Two typical frequency-multiplier chains.

the signal applied to the grid is of $V2$ negative and $V2$ remains cut off. On the next half-cycle of input voltage, $V2$ conducts and $V1$ is cut off. The plates of $V1$ and $V2$ are connected in parallel; therefore two pulses excite the tank circuit for each cycle of input. These pulses therefore drive the tank at a frequency twice that of the input. The output can be compared to the ripple present in the output of the unfiltered full-wave rectifier circuit. Because of its simplicity, this circuit is widely used in f-m transmitters. If compact design is desired, the circuit can serve also as combination doubler and power amplifier because of its relatively high efficiency and low output of undesirable harmonics.

e. Multiplier Operation at High Frequencies. The operation of frequency multipliers at high frequencies is hindered by degenerative effects tending to lower the power output. These effects can be caused by capacitive or inductive feedback. Capacitive coupling between grid and plate circuits caused by the interelectrode capacitance gives rise to degeneration which is equivalent to loading the output circuit. It has

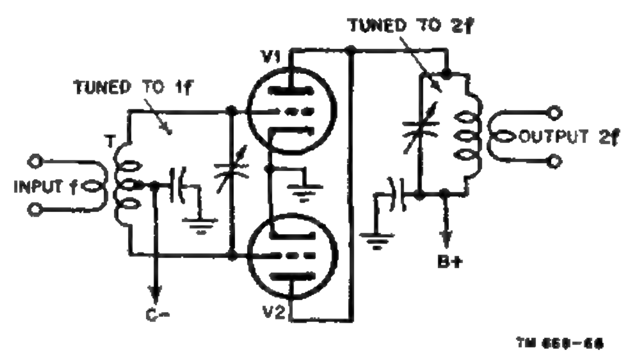


Figure 66. Push-push doubler.

been found that power tubes with large values of mutual conductance and grid-plate capacitance are rendered inoperative as multipliers at high frequencies by this loading effect. Before the circuit can be made to operate with full efficiency and power output this loading effect must be *neutralized.* Three methods of canceling the effects of the grid-plate capacitance are illustrated in figure 67.

(1) The circuit in A shows one method of neutralizing the grid-plate capacitance, C_{gp}. Capacitor C_n provides feed-

back which is 180° out of phase with the degenerative feedback caused by C_{gp}. By selecting the correct value of C_n, the two feedbacks are made equal, but out of phase, and cancel. Care must be taken not to make C_n too large, or the regenerative feedback will cause oscillation. C_n usually is adjusted for minimum d-c plate current with the plate circuit at resonance. Another method, shown in B, is to insert from plate to grid a series *L-C* circuit which is antiresonant at both input and output frequencies. This has the effect of placing a high-impedance feedback path from plate to grid. C_n is adjusted for minimum plate current with the circuit at resonance. If the input circuit between grid and cathode presents a low inductive reactance to the feedback voltage, it will serve to balance out the effects of degenerative feedback. By inserting the proper amount of inductance, as shown in C, in series with the grid circuit, the degenerative feedback is canceled. L_n must not be made too large, or the circuit will oscillate. In a v-h-f (very-high-frequency) multiplier, lengthening the grid lead may be sufficient to provide this inductance.

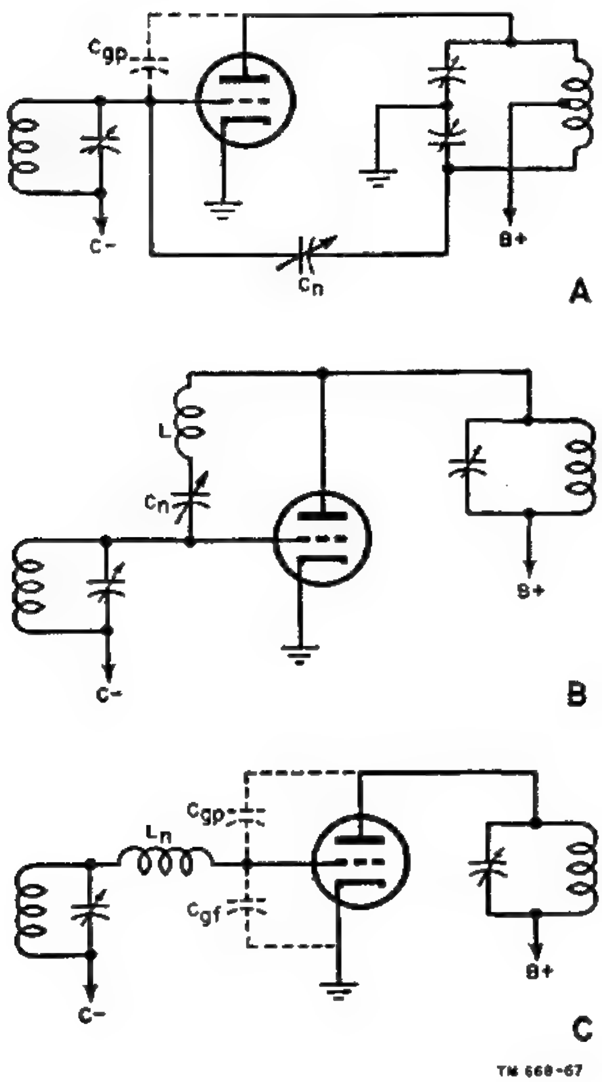


Figure 67. Neutralizing circuits.

(2) Degeneration equivalent to an output loading effect also can be produced by cathode inductance common to both plate and cathode circuits. This inductance prevents the cathode from being connected directly to ground, and plate current flowing through the cathode circuit creates a reactive voltage which opposes the input signal. To obtain efficient operation at very-high frequencies, the cathode inductance must be neutralized at both input and output frequencies. A method for accomplishing this is illustrated in figure 68. The circuit is a series combination of $L1$ and $C1$ shunted by capacitor $C2$, the combination being in series with the cathode lead. $L1$ and $C1$ in series with L_k form a circuit providing series resonance at the input frequency. L_k and shunt capacitor $C2$ form a series resonant circuit to ground at the output frequency.

(3) Several tetrodes and pentodes have been designed especially for use at very-high frequencies. In these, two separate leads for grid and plate returns to the cathodes have the effect of reducing degeneration. When high output and efficiency are needed, these tubes can be used in a push-pull arrangement. Also, if the screen grid of a tetrode or pentode is effective at the frequency of operation, no neutralization is required.

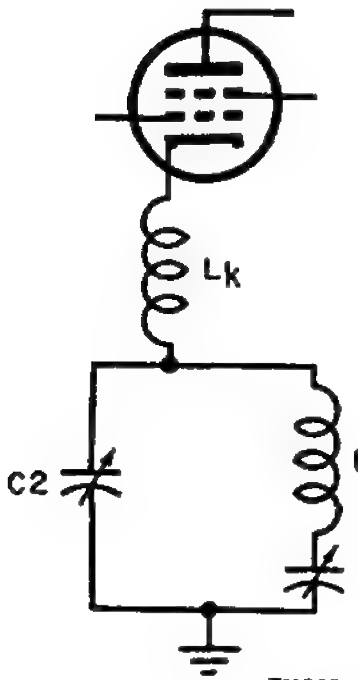


Figure 68. Cathode inductance neutralization.

f. Other Frequency Multipliers.

(1) Mixers similar to those in superheterodyne receivers can be used for frequency multiplication. However, they give no increase in effective deviation. It is possible also to synchronize an oscillator running at a higher frequency with one at a lower frequency, if the two frequencies are multiples of one another. The synchronized oscillator is used more frequently, however, as a frequency *divider.*

(2) Higher multiples of a given frequency can be obtained by using a nonlinear device that produces harmonics. The distortion of the grid current in a class C amplifier is one method. A second method produces harmonics in a mixer or other nonlinear modulator. The desired harmonic is amplified and then fed back to the mixer, where it reinforces the output at its own frequency. This device is called a regenerative modulator. Since subharmonics can be selected as well as harmonics, the device can be used also as a frequency divider. In practice, this has been its principal application.

g. Combined Frequency Multiplier, Master Oscillator.

(1) The master oscillator and a multiplier can be combined in a circuit using only one tube. In figure 69, such a circuit combines the oscillator and multiplier in a single pentode tube. Elimination of components and reduced current drain are gained at the expense of only a slight loss in oscillator stability.

(2) The oscillator is a simple Colpitts, with r-f oscillations generated in the control-grid, screen-grid circuit of the tube. The output frequency is selected by a tuned plate load, *L*2-*C*2. The tank circuit, *L*1-*C*1, in conjunction with *C*8 and *C*9 forms the fundamental frequency-determining components. The grid-leak bias for the operation of the oscillator is provided by *R*1 and *C*4. The r-f choke in the cathode circuit permits d-c current to re-

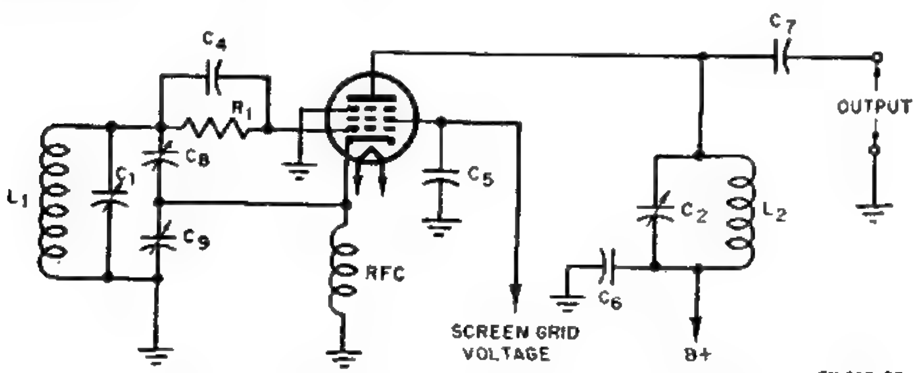


Figure 69. Typical combined oscillator, frequency multiplier.

turn to the negative side of the power supply, at the same time maintaining the r-f potential at the cathode. The screen is effectively bypassed for r-f by $C5$. Therefore, the screen can act as a grounded plate, with the control grid and cathode serving their normal functions. The result is a triode oscillator circuit.

(3) The current pulses generated by the oscillator section reach the plate flow through the tuned resonant circuit formed by $L2$ and $C2$. This circuit presents a high impedance to the harmonic frequency of the plate current pulses since it is tuned to resonance with it. Consequently, a considerable harmonic voltage is developed between plate and ground. The lower part of the plate tank circuit is bypassed to ground through capacitor $C6$ and the output from the stage is coupled capacitively to the following stage through $C7$.

(4) Since the output circuit is coupled to the oscillator circuit through the electron stream alone, there is comparatively little interaction. If the screen voltage is set properly, it is possible to reduce the variation in operating frequency with changes in tuning of the output circuit to a low value. The higher the order of harmonic to which the plate circuit is tuned, the better the stability of the oscillator. Any interaction between output and oscillator circuits must come as a result of Miller effect between the two circuits. Capacitive coupling between the grid and plate, which tends to cause this interaction, is considerably reduced by the shielding effect of the screen grid. Therefore, the grounding of this grid through capacitor $C5$ must be complete to obtain maximum isolation.

(5) The circuit of figure 69 is one possible way in which a frequency multiplier can be combined directly with the oscillator. Any oscillator circuit that can operate with its plate at ground potential can be substituted for the Colpitts circuit shown. The frequency-multiplier action is the same regardless of the oscillator, the only advantage gained with any specific circuit being attributable to the characteristics of the oscillator itself.

41. Power Amplifiers

a. F-M Power Amplifiers.

(1) The requirements for an f-m power amplifier are somewhat different from those for a-m, in which the power amplifier is usually the stage in which the modulation is introduced. Therefore, any losses that take place during the modulation process must be dissipated in the power amplifier stage. Since the f-m power amplifier has no connection with the modulation process, the only losses that are involved are those inherent in the tube and circuit when amplifying an unmodulated carrier.

(2) When a-m is produced in a low-level stage, it is necessary that the power amplifiers reproduce the modulation envelope without distortion, and therefore linear amplifiers must be used. F-m, which is also produced at a low level, does not have a modulation envelope that can be distorted by the limiting action of highly efficient class C amplifiers. The characteristics of an f-m power amplifier are determined at class C, c-w ratings.

b. Class C Amplifiers.

(1) A typical class C amplifier, as used for f-m signals, is shown in figure 70. The input signal is supplied through a tuned transformer, $T1$. Output is developed by the r-f signal appearing across the parallel-resonant circuit, $T2$.

(2) Figure 71 shows the relationship between the various voltages and currents in the circuit of figure 70. The grid bias, e_c, is developed across the tuned circuit of $T1$. The class C amplifier operates with grid bias much greater than cut-off. Therefore, the

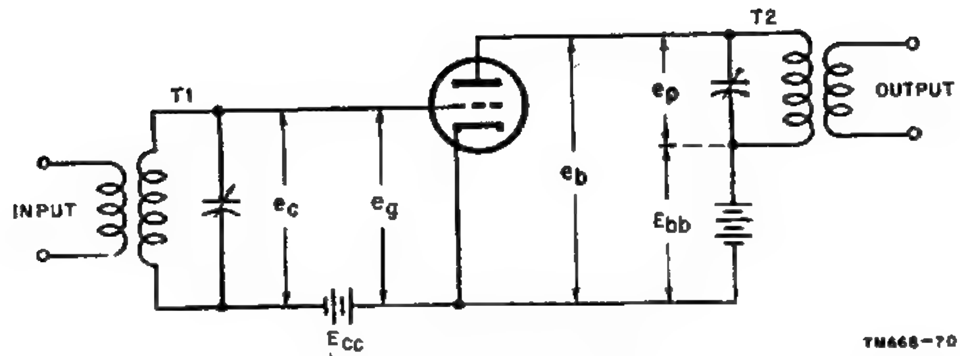


Figure 70. Class C amplifier.

grid excitation voltage causes plate current to flow during only part of the cycle. During the remainder of the cycle, the voltage on the grid is below the cut-off value, the plate current, i_b, is zero, and the corresponding plate voltage, e_b, rises to its highest value, or E_{bb}. Since no plate current flows, the voltage drop across the plate load impedance must be zero. The voltage drop across the load, therefore, is 180° out of phase with the grid voltage. The a-c components of the plate and grid voltages are sinusoidal because of the sharply tuned resonant circuits.

(3) Plate current flows when the grid voltage, e_c, rises above cut-off. The angle of flow of plate current is ϕ_p and is usually less than half a cycle. Grid current flows during the angle ϕ_g when the grid voltage, e_c, becomes positive. The sum of these two currents, $i_b + i_c$, is the space current, i_s, and represents the total current leaving the cathode. The angle of grid current flow depends on the *ratio* of the *grid bias* to the *peak signal amplitude.* This is equivalent to saying that, in a particular amplifier, the value of the grid bias chosen determines the angle of plate current flow for a given input signal. Short angles of flow give high efficiency and low power output, whereas large angles give low efficiency and higher power output.

(4) At any moment, the total power input to the plate is the product of the total voltage, e_b, supplied to the plate, and the instantaneous plate current, i_b. The power output is equal to the product of the load voltage and the plate current. The power loss at the plate is the difference between the input power and the output power. The efficiency of a class C amplifier is the ratio in percent of the output to input power and is usually between 60 and 80 percent. This high efficiency is possible because the plate current flows only when most of the voltage drop is across the output circuit. Therefore, only a small part of the supply voltage is wasted as a voltage drop between the plate and cathode of the tube.

(5) Since the grid of the tube swings positive and draws current during part of the cycle, power is absorbed from the excitation circuit, which is the product of the exciting voltage, e_c, and the grid current, i_c. Some of this power is lost at the grid, and the remainder is dissipated in the bias battery. If grid-leak bias is used, the remainder is dissipated as heat in the grid-leak resistor.

c. Class C Power Amplifiers With F-M Excitation.

(1) An f-m wave will not be distorted in passing through such an amplifier, since the frequency of the voltage developed at the output is the same as that provided by the grid excitation. If the input signal deviates through

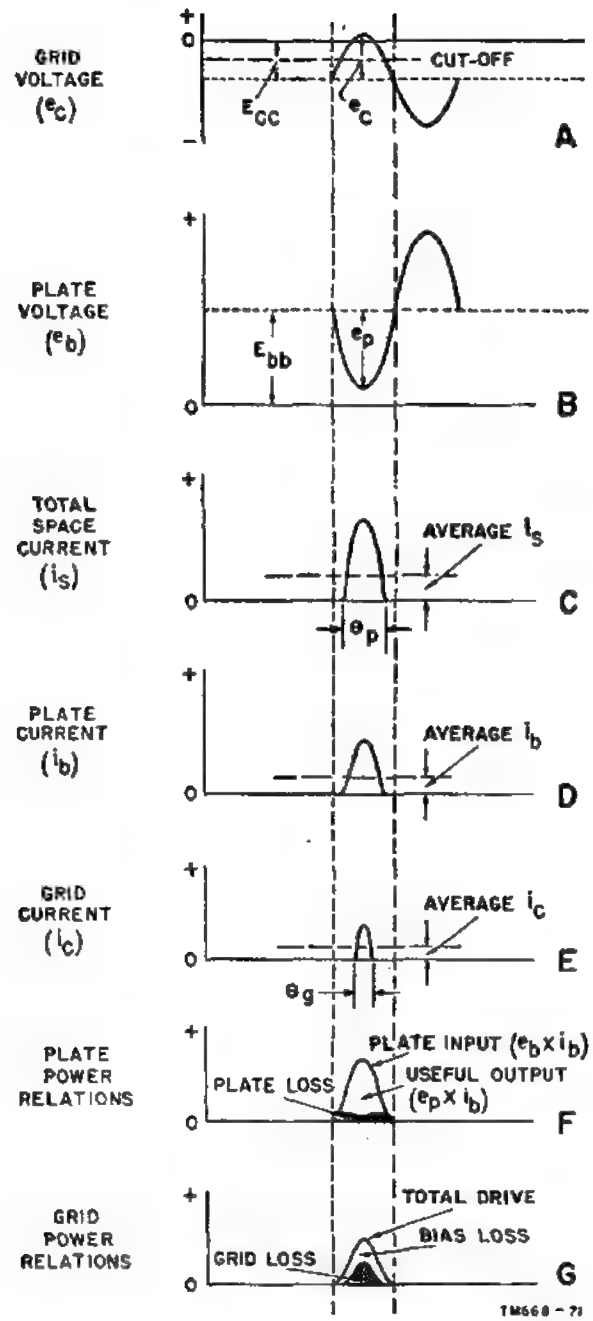


Figure 71. Current and voltage in Class C amplifiers.

a number of cycles, the output signal deviates by the same amount. Since resonant circuits are used in both input and output circuits, it is necessary that they have broad enough selectivity that the full frequency band of the modulated carrier is passed without reducing the amplitudes of the outer sidebands. This is possible only where the total bandwidth of the f-m signal is not large compared with the carrier frequency. This places a restriction on the maximum usable deviation.

(2) An examination of the circuit in figure 70 shows that, since the grid and plate circuits are tuned to the same frequency, energy can feed back through the grid-plate capacitance, permitting tuned-plate, tuned-grid oscillation at the frequency of the tuned circuits. In all triode amplifiers, this tendency toward oscillation must be neutralized. Figure 72 shows the same circuit as figure 70, with the addition of a tapped plate-tank coil grounded at the center by a bypass capacitor. From the side of the coil opposite the plate connection, a small variable capacitor, C_n, is connected to the grid. Since the opposite sides of the coil are 180° out of phase, the voltage tapped by the small variable capacitor is therefore 180° out of phase with the plate voltage. It is also out of phase with any grid voltage fed back through the grid-plate capacitance. The variable capacitor, along with the interelectrode capacitance from grid to cathode (shown in dashed lines), acts as an adjustable voltage divider which permits a variable amount of out-of-phase voltage to be applied to the grid. The voltage tending to cause oscillation is out of phase with this neutralizing voltage. It therefore can be canceled out if the two are made equal by proper adjustment of the neutralizing capacitor. The neutralizing capacitor is of approximately the same size as the grid-plate capacitance. It usually is made slightly larger in practical high-frequency circuits. This is done because inductance in the connecting leads produces reactance opposite to that of the capacitor and tends to reduce its effectiveness.

d. Class C Tetrode Amplifiers.

(1) The disadvantages of ordinary triode amplifiers in the frequency range

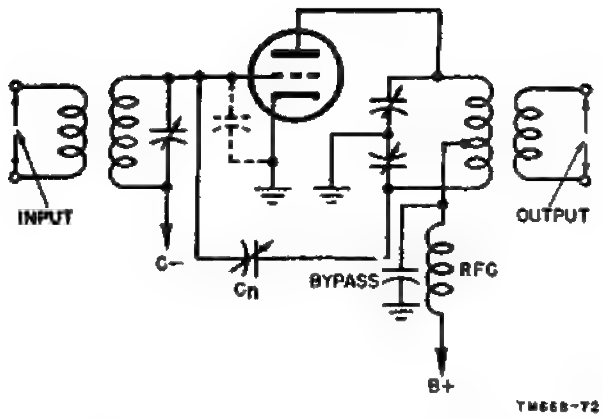


Figure 72. Neutralized triode amplifier.

where f-m usually operates are many. The necessity for neutralization means that an additional adjustment is needed and, as the frequency increases, this becomes more and more critical. Because of their low power sensitivity and high output capacitance, a large amount of grid excitation power is needed to produce a given amount of output power and the excitation requirements increase as the frequency increases. The high output capacitance of triodes also reduces the value of usable tank-circuit inductance, resulting in high Q and too narrow a band pass. At higher frequencies, where it is important to conserve the number of tubes and the amount of total power input, tetrodes are more useful.

(2) The operation of a tetrode amplifier for the very-high frequencies used for f-m transmitters is somewhat different from that for the triode. The grid bias is set with reference to screen-current cut-off, as compared to plate-current cut-off in a triode. The angle of flow therefore depends largely on the screen-grid voltage and control-grid bias. The maximum grid voltage must not be greater than the screen voltage. In addition, the minimum plate voltage during the operating cycle must not be less than the screen voltage (except in beam tetrodes). If this last condition occurs, the plate emits secondary electrons that are collected by the screen, making the plate a virtual cathode. The space charge in a beam tetrode (or the suppressor in a pentode) prevents this effect.

(3) In the higher part of the v-h-f range, it is no longer possible to use many tetrodes as class C amplifiers for f-m. At these frequencies, the inductance of lead wires and internal tube supports produces enough reactance to prevent the screen from being effectively grounded to r-f currents. Therefore oscillation can occur. For limited frequency ranges, it is possible to reduce the impedance from screen to ground by making the screen bypass capacitor and the screen lead inductance a series resonant circuit. However, if the circuit must be used over a wide frequency range, which is common with f-m equipment, an additional screen neutralizing control must be added, the operation of which is very critical. To overcome this difficulty, special circuits must be used.

e. Grounded-Grid Triode Amplifiers.

(1) There are three types of triode amplifiers, the type depending on the manner in which the signal is applied to obtain an output. The load may be connected between the plate and cathode, with the signal applied between the grid and cathode. If the common element is placed at zero potential, the stage is called a *grounded-cathode* amplifier. In the *cathode follower*, the plate is grounded to r-f, the signal is applied between grid and ground, and the load is placed between the cathode and ground. Finally, in the *grounded grid amplifier* (fig. 73), the signal is applied between the cathode and ground, the grid is grounded, and the output is taken across a load between plate and ground.

(2) The grounded-grid circuit permits a triode to be operated at high frequen-

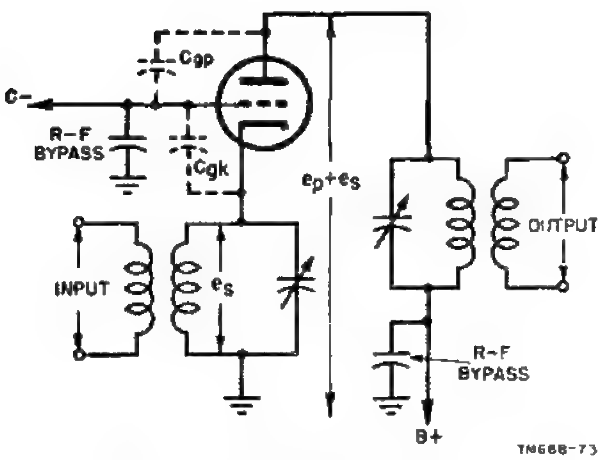


Figure 73. Grounded-grid amplifier.

cies without neutralization. Therefore, one of the most objectionable features of a triode r-f power amplifier is overcome. In this circuit, the grid is grounded through an r-f bypass capacitor and serves as a shield between the input and output circuits, thus preventing feedback of energy and resultant oscillation. It also has the advantage of very low output capacitance, since the only capacitance across the output added by the tube is that between grid and plate (fig. 73). In tubes designed especially for this purpose, the capacitance is made very low and larger values of inductance can be used in the plate circuit at relatively high frequencies. This results in higher efficiency.

(3) Another characteristic feature of the grounded-grid amplifier is that both the driver stage, which supplies the input, and the amplifier stage itself supply the plate load circuit. Note that the driver produces an r-f voltage e_s across the input terminals. An r-f voltage also is produced across the plate and cathode elements of the tube. These voltages are 180° out of phase in respect to the cathode, and therefore the r-f output voltage from plate to ground is the sum of the two out-of-phase voltages.

(4) The plate current generally is 180° out of phase with the plate voltage. This means that the signal current flowing in the cathode circuit must be the same as the plate current. The cathode can have low impedance, and the plate circuit can have high impedance. Therefore, the tube acts as a device to transfer the space current from a low impedance to a high impedance. The output power is proportional to the square of the current multiplied by the resistance; therefore, the input (cathode power) is low, the output (plate power) is high, and the tube acts as a power amplifier. The gain of the amplifier is proportional to the ratio of the output impedance to the input impedance.

(5) Because the input impedance can be made small, the bandwidth of the input circuit can be very great. Since the output capacitance is only that from plate to grid, the inductance in the plate circuit can be made large for a given resonant frequency. Therefore, the selectivity of the output circuit also can be made broad.

f. Push-Pull Amplifiers for F-M.

(1) The grounded grid push-pull amplifier is used frequently in high-power f-m systems at very-high frequencies. This circuit can be used also in the microwave region with specially constructed tubes and circuits. When two tubes are within one envelope, degeneration resulting from the inductance of the cathode lead is canceled out. This permits the use of lower values of grid drive than would be necessary with two separate tubes or with a single-ended stage. Typical circuits for push-pull triode and beam-tetrode amplifiers are shown in figure 74. Because two tubes are involved, twice the amount of excitation must be supplied. However, the output capacitance placed across the plate tank circuit and the input capacitance across the grid tank are halved because the tube capacitances are in series, and therefore equal to only half the value of one alone.

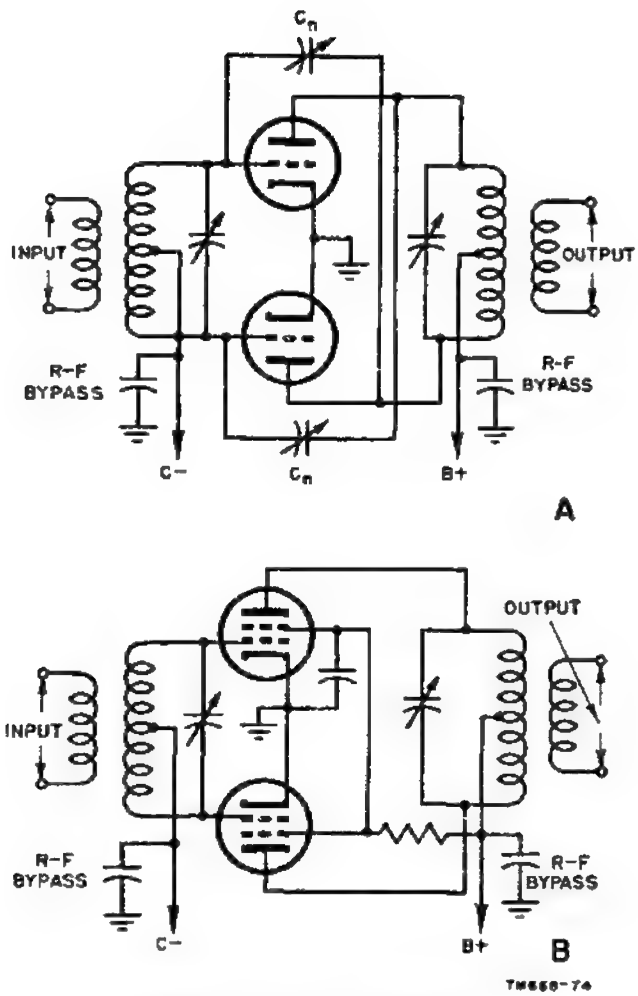


Figure 74. Push-pull amplifiers.

(2) At high frequencies, the necessity for neutralization of the push-pull triode amplifiers with its attendant difficulties makes the use of tetrodes desirable. These are combined in one envelope for high frequencies so that the internal inductance of the leads and tube elements does not interfere with operation. The usual tube construction has a single cathode which eliminates the problem of separate cathode-lead inductance. Neutralization of any amplifier increases the output capacitance of the tube. The push-pull cross-neutralized amplifier shown in A of figure 74 has an output capacitance equal to the plate-cathode capacitance of each tube in series plus twice the grid-plate capacitance. This extra capacitance is added by the neutralizing circuit. It limits the operating frequency of the amplifier because the output capacitance is a major factor in determining the plate-tank constants at very high frequencies.

(3) Some of these difficulties can be overcome by using push-pull, grounded-grid amplifiers. The operation of these amplifiers is similar to the single-ended stage, the major difference being the change in the cathode input impedance. Both cathode loads are effectively connected in series, and the input impedance becomes four times the value of one tube alone. Each part of the cathode load acts independently for its associated tube and therefore the voltage across both loads is twice that of the individual load. When the voltage across the secondary is doubled, the impedance, which is proportional to the square of the voltage, is increased by a factor of four. The same thing is true of the output load impedance in all push-pull amplifiers. The voltage across each section of the primary is the same as for a single-ended amplifier. The doubled voltage of the push-pull connection requires a total load impedance of four times the impedance for one side of the load. This makes the requirements for tank-circuit inductance easier to meet at very-high frequencies and accounts for the wide use of push-pull circuits for f-m power amplifiers.

g. Power-Amplifier Input Circuits.

(1) It is highly desirable to have as efficient a transfer of power from the driver stage to the power amplifier as possible. Therefore, the grid tank circuit must provide an impedance match between the grid input impedance of the amplifier and the plate output impedance of the driver stage. If a grounded-grid amplifier is used, similar considerations apply to the cathode tank circuit.

(2) The impedance of a circuit normally is defined as *the ratio of voltage to current.* However, in the grid circuit of a class C amplifier, this ratio is far from constant. When the grid voltage goes highly negative, no current is drawn at all; when it is positive, a great deal of current flows. Therefore, the impedance of the grid circuit varies over a range from an extremely high to an extremely low value through the operating cycle. If the *input impedance of the grid circuit* is too high, the heavy current demanded by the extreme positive grid swing cannot be drawn. As a result, actual grid voltage and consequent loss of peak efficiency are reduced in the operation of the amplifier. If the impedance of the grid tank circuit is too low, a great deal of power from the driver stage is required to operate it, and the losses in the inductor consume a considerable amount of the applied power. Generally, a compromise value is used which is approximately equal to the ratio of the driving power in watts divided by the square of the grid current. The choice of values for the components in the grid tank circuit is determined by this impedance. The result usually is satisfactory regulation of the grid voltage without excessive power loss.

(3) Some of the actual circuits used at the grids of single-ended, push-pull, and grounded-grid amplifiers are shown in figure 75. The simplest of the capacitance-coupled, tuned-input circuits for a single-ended power amplifier is shown in A. The inductor, *L*, and the capacitor, *C*, constitute a tank circuit. The inductor is tapped so that it steps up the signal voltage from the plate circuit of the driver tube. The signal is coupled from the top of the plate load to the grid through C_c. Bias is supplied to the grid through an r-f choke. The circuit of B is much the same as that of A, except that the tuned circuit is now in the grid circuit of the power amplifier, and driver plate voltage is supplied through an r-f choke. This arrangement is used when the required impedance at the grid of the power amplifier is lower than the output impedance needed in the driver stage.

(4) The circuit in C permits complete d-c isolation of the tuned circuit from the driver and amplifier stages. Capacitors *C*1 and *C*2 block the d-c voltages and at the same time couple the signal from driver plate to amplifier grid. Bias is supplied through an r-f choke. This circuit provides no means for adjusting the impedance between the grid and plate circuits.

(5) The circuit of D permits the driver stage to be neutralized. It also provides variable drive for the amplifier grid and d-c isolation for the bias and high-voltage circuits without the need for r-f chokes. The inductor, *L*, of the tuned circuit is split into two parts at the center, each of which is grounded separately with r-f bypass capacitors. High voltage for the driver plate therefore cannot reach the grid of the amplifier, and bias can be applied at the center tap as shown. Neutralization for the driver is obtained from the side of the tuned circuit opposite the plate through C_n.

(6) The capacitively coupled circuit in E also permits neutralization of the driver. Driver plate voltage is fed to the center tap of *L* through an r-f choke. A split-stator tuning capacitor, *C*, is used to tune the circuit to resonance. The neutralizing voltage is derived from the lower end of the coil and fed back through C_n as before; the grid drive for the amplifier is coupled through capacitor C_c from the same point. Bias for the amplifier grid is introduced through the r-f choke, as shown.

(7) In F, the tuned-plate tank of the driver is inductively coupled through a low impedance link to a tuned-grid

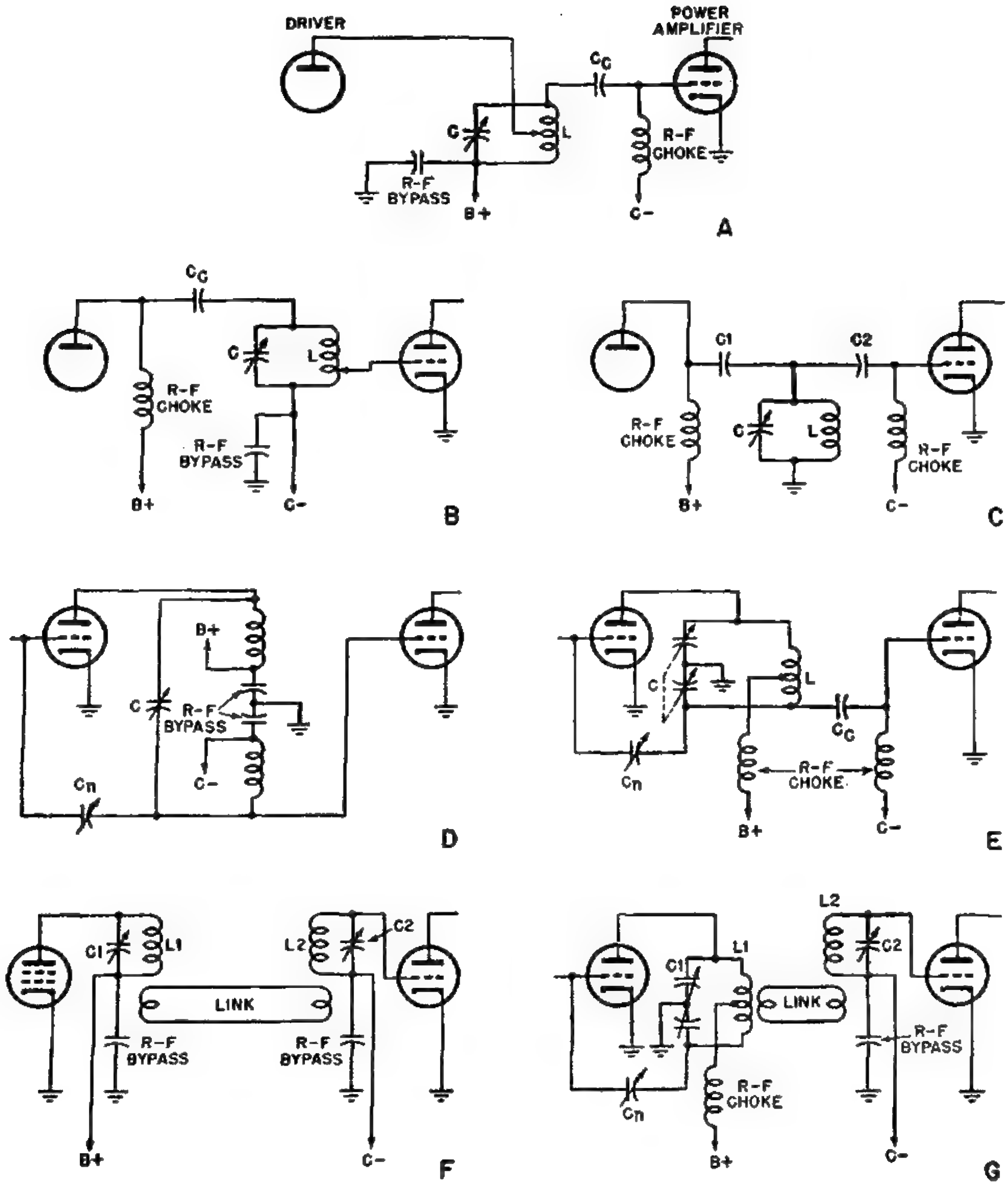


Figure 75. Practical power amplifier grid tank circuits.

circuit. The link inductance is small, and therefore the impedance of the coupling circuit is low. This minimizes losses in the transmission of driving power, and at the same time provides great flexibility in matching impedances between the driver and the power amplifier. The equivalent arrangement used with a neutralized driver is shown in G. The high voltage is applied to the tuned circuit of $L1$ and $C1$ of the driver and the neutralization is accomplished by C_n. The link circuit is slightly different in that the coupling link from the driver is positioned in the center of $L1$ rather than at its lower end.

(8) The input circuits for push-pull amplifiers (fig. 76) are variations of those used in single-ended amplifiers. The capacitive coupling arrangement in A uses a tuned-plate circuit for the driver tube formed by $L1$ and $C1$. Because each half of the split coil is out of phase with the other half, each half can supply grid drive in push-pull directly through coupling capacitors $C2$ and $C3$. A split-stator capacitor, $C1$, is used as the main tuning capacitor for the coupling arrangement. Two r-f chokes are used, one from each grid to the bias supply. Capacitor $C4$ introduces a small amount of capacitance from the lower end of the tank circuit to ground in order to compensate for the plate-to-ground capacitance of the driver tube, which appears across the upper half of the circuit. Driver plate voltage is applied through a suitable r-f choke. Neutralization of the driver stage is provided by C_n.

(9) An inductive coupling arrangement for the grid tank circuit is shown in B. The link circuit transfers energy from the driver tank, $L1$-$C1$; the out-of-phase voltages to the push-pull grids of the amplifier are developed across the split-stator capacitor, $C2$. Bias is introduced through an r-f choke. In C, the coil is center-tapped and grounded by a capacitor, and grid bias is introduced directly at the center tap. A third variation of the circuit is shown in D, where neither the split-stator capacitor nor the coil is grounded directly. $C2$ is grounded through an r-f bypass capacitor. The resistor, R, is low in value and is placed in series with the center tap of coil $L2$ and tuning capacitor $C2$. It serves to equalize any slight variation in the center tap of the coil. Bias is supplied at the center tap on $L2$.

(10) The choice of one or another of the circuits mentioned above depends on several considerations. Where the tuning capacitor is grounded directly, it must have twice the voltage rating of one that is grounded through a capacitor. In some instances, the use of an r-f choke for bias is undesirable; therefore, one of the capacitor arrangements must be used. The choice of inductive or capacitive coupling depends on the character of impedance matching and therefore indirectly on the frequency of operation. Generally, at the higher frequencies, the simpler coupling arrangements are preferred because of their lower losses. Finally, the link-coupled circuits permit the driver to be located at some distance from the amplifier and connected to it through a low-impedance transmission line.

h. Power-Amplifier Output Coupling Networks.

(1) All power amplifiers used at high frequencies must be coupled to a load circuit. Generally the impedance of the load is not the same as the plate-circuit load required by the amplifier. At the very-high frequencies where f-m is used, the amplifier is coupled to the antenna through a transmission line. In small portable units, the antenna usually is connected to the amplifier tank circuit by a different type of coupling arrangement. The impedance of the load must be matched to

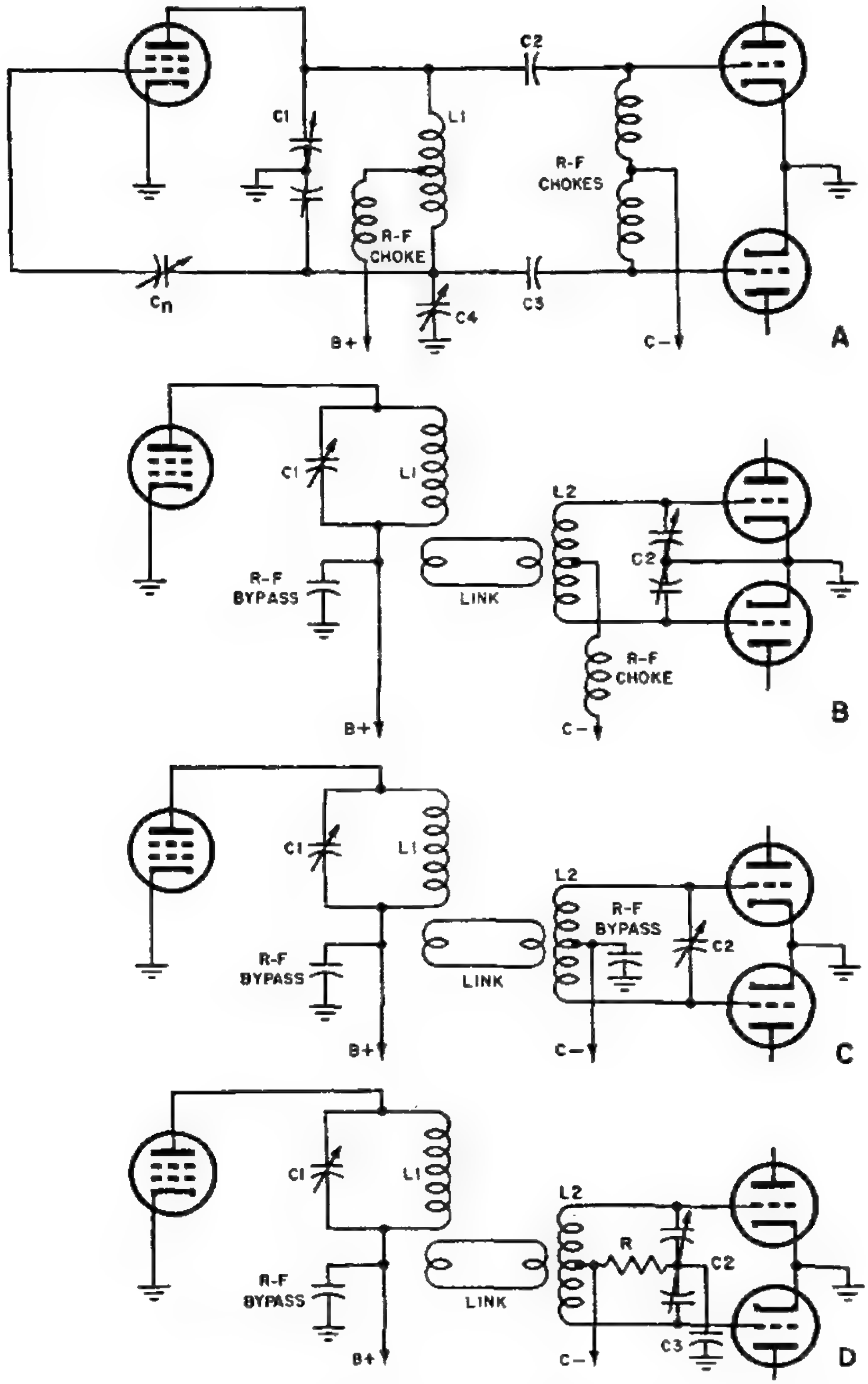


Figure 76. Grid tank circuits for push-pull amplifiers.

the plate impedance of the tube for maximum transfer of energy. There are many practical means of accomplishing this, and a variety of inductive and capacitive coupling circuits are used.

(2) The effect of the coupling network on the power amplifier is much the same regardless of the particular network involved. Analysis of the circuit shown in A of figure 77, shows that the output of the amplifier developed across the high-impedance plate tank circuit load is coupled to a low-impedance transmission line by a tuned step-down transformer. The circuit must perform the following functions:

(a) The input impedance of the coupling network must meet the tube requirements for load impedance.

(b) The transfer efficiency of the network (the ratio of output power to input power) should be as high as possible.

(c) The input impedance at all frequencies except that to be amplified should be small so that spurious frequencies generated in the amplifier are minimized.

(d) The selectivity of the network should suppress unwanted frequencies and prevent their transfer to the load. The selectivity must not be so high that outer side bands of the frequency-modulated signal are reduced in amplitude.

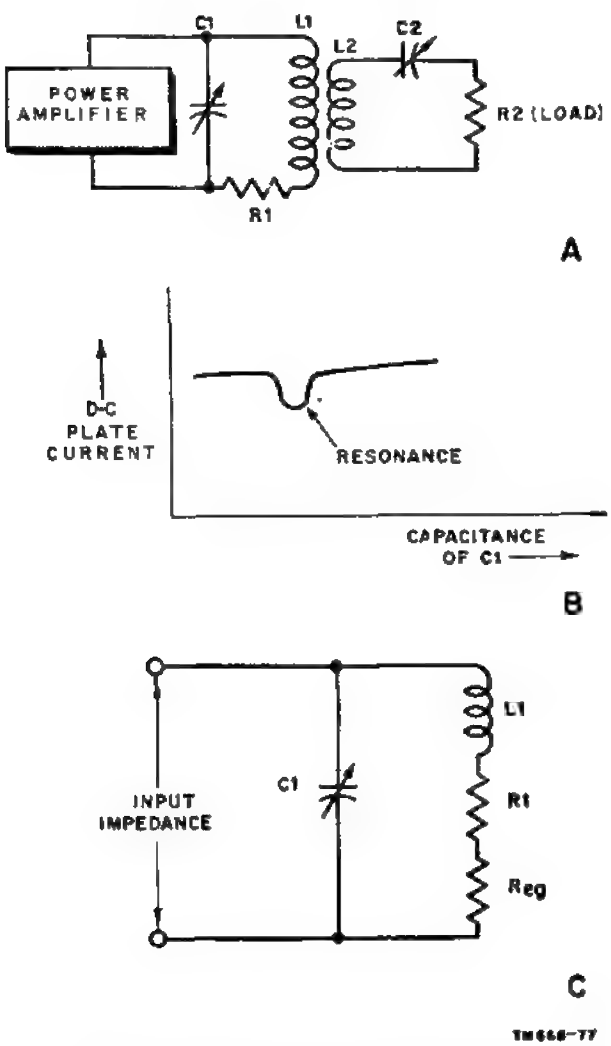


Figure 77. Ideal power amplifier output coupling network.

(3) Capacitor $C2$ in A is adjusted so that it resonates with $L2$ at the center operating frequency. The coupling between $L1$ and $L2$ then is adjusted to present approximately the desired load impedance to the tube. Finally, $C1$ is varied until the circuit presents a purely resistive impedance to the tube. At this point, the d-c plate current of the amplifier is a minimum because the power factor is unity. The tuning curve of d-c plate current versus capacitance is shown in B.

(4) The load impedance presented to an amplifier by a parallel circuit at resonance is equal to the product of the operating Q and the inductive reactance. The actual Q of the inductance is many times higher than the operating Q since the load that is reflected into the parallel circuit by transformer action reduces the effective Q of the circuit. This is equivalent to placing additional resistance R_{eq} in series with the actual resistance, $R1$, of coil $L1$, as shown in the equivalent circuit of C. The effect of coupling a load to a parallel-resonant circuit is to decrease the effective Q. For high efficiency,

the unloaded Q of the tank circuit should be as high as possible. The loaded Q then should be made as low as possible. However, this conflicts with the requirements of selectivity and low impedance to spurious frequencies. Therefore, a compromise value is chosen. Typical values involve unloaded Q's of 200 and loaded Q's of 10 to 12 in v-h-f f-m transmitters. The transmitter efficiency of the tank circuit can be shown to be as follows:

$$\text{Transfer efficiency} = \frac{Q_o - Q_L}{Q_o} \times 100 \text{ percent}$$

where Q_o is the unloaded Q of the parallel-resonant circuit and Q_L is the loaded Q. For the values given above, the efficiency is

$$\frac{200-12}{200} \times 100 = 94 \text{ percent}$$

i. Practical Transmitter Inductively Coupled Tank Circuits.

(1) A large variety of practical circuits have been devised which present the proper load impedance to the power amplifier when connected to the transmission line or antenna. Some of the coupling circuits that have not been discussed are illustrated in figure 78. The simple, parallel-resonant tuned circuit in A frequently is used for single-ended tetrode amplifiers. The circuit is shunt-fed, the plate voltage being applied in parallel with the tank circuit. The d-c plate voltage applied through the inductance is fed through an r-f choke which effectively isolates the power supply. The tank circuit, $C1$-$L1$, is coupled to the plate by capacitor $C2$. The advantage of this circuit lies in the removal of all d-c voltages from the tuning capacitor. This means a lower value of total voltage across this component, with correspondingly smaller size and lighter weight. A major shock hazard from contact with an exposed portion of the tank circuit is removed. However, the possibility of a bad r-f burn always exists.

(2) Although the circuit in A of figure 78 is satisfactory when used with tet-

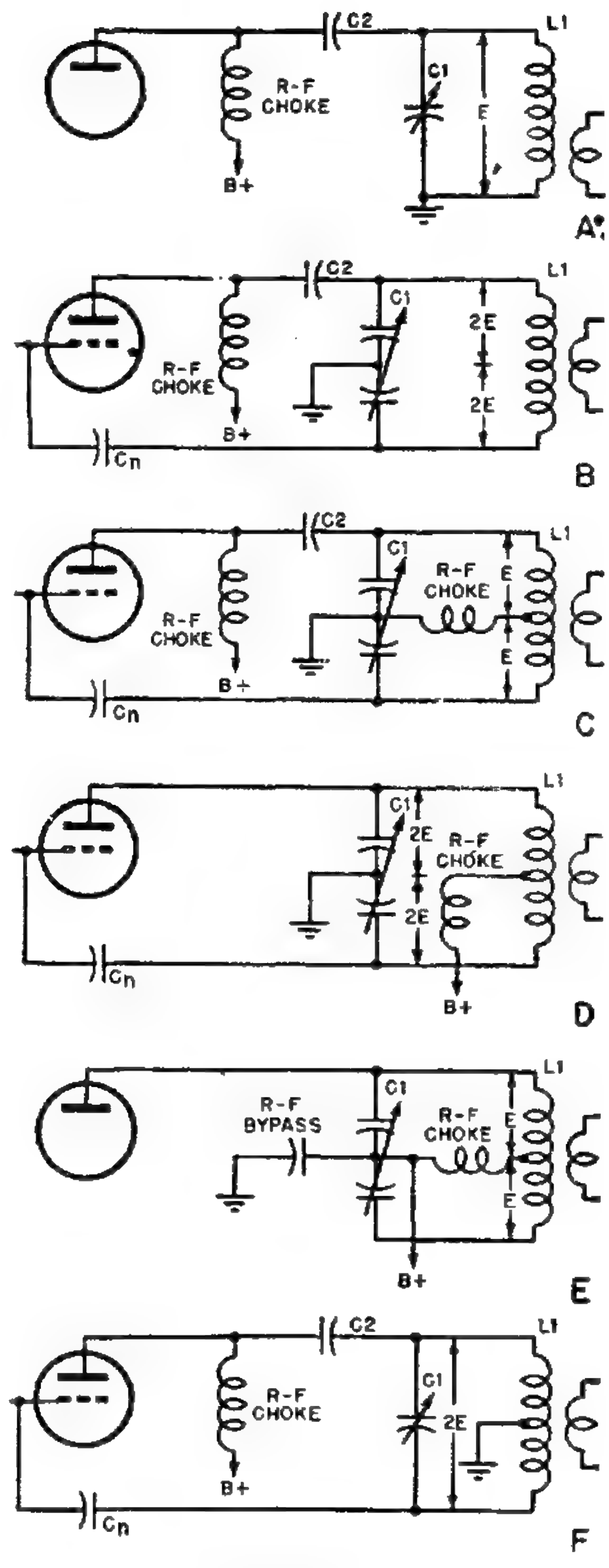


Figure 78. Practical plate-tank circuits for class-C amplifiers.

rodes and grounded-grid triodes, it provides no means of neutralizing an ordinary grounded-cathode triode. In B, the capacitor C, has a split stator with the rotor directly grounded. This permits an out-of-phase voltage to be taken from the lower end of the coil and returned to the input through the neutralizing capacitor, C_n. Plate voltage is applied as before through an r-f choke. The blocking capacitor, $C2$, prevents the plate voltage from reaching the tuned circuit. However, the split-stator capacitor effectively divides the circuit in two parts, and an r-f peak of twice the d-c plate voltage can appear across each. This requires a physically large capacitor and limits the use of this circuit.

(3) The problem of excessive capacitor voltage is solved in the circuit in C, where shunt feed is retained along with the neutralizing circuit. However, the coil is center-tapped and a small r-f choke places the center of the coil at the d-c ground potential. The over-all voltage across each half of the circuit becomes that of the a-c voltage alone, and permits the size of the tuning capacitor plate spacing to be reduced.

(4) Another tank circuit which allows the neutralization of a single-ended stage is shown in D. This circuit is series-fed through an r-f choke to the center of the tapped tank coil. A split-stator capacitor is used for tuning and the neutralizing voltage is taken from the lower end of the coil through C_n. This circuit has the same faults as the one in B, since the r-f equivalent of twice the plate voltage appears across each half of the circuit. It can be remedied by the circuit of E, where the rotor of the tuning capacitor is left ungrounded for d-c, but is bypassed for r-f by a capacitor. The power-supply voltage is applied to the rotor to equalize the voltages that would otherwise be built up across the tank. This results in a severe operating hazard unless the shaft of the tuning capacitor is well insulated from the tuning knob.

(5) The circuit at F is a modification of this arrangement, permitting the use of shunt feed and doing away with the necessity for a split-stator capacitor. Plate voltage is applied through an r-f choke, and the d-c voltage is prevented from reaching the tank by a blocking capacitor, $C2$. The tank inductance itself is grounded directly at the center tap, making this circuit desirable when a grounded coil is needed. This is the case where the coils are selected by a rotary indexing switch when changing frequency of operation over a wide range.

(6) Almost all of the preceding circuits for single-ended stages have their push pull counterparts. Some of these are shown in figure 79. The circuit in A is the push-pull counterpart of the simple resonant tank. A split-stator capacitor is used with the push-pull version, and the rotor is grounded for r-f through a bypass capacitor. Plate voltage is series-fed to the center of the inductor through an r-f choke. To reduce the voltage across each half of the tuning capacitor, the plate voltage sometimes is connected to the rotor. The shock hazard introduced by this can be avoided by grounding the rotor of the tuning capacitor directly, and applying the plate voltage through an r-f choke, as in B. This circuit has an r-f peak of twice the d-c plate voltage appearing across each section of the capacitor.

(7) A third alternative, which is less desirable than the other two, is shown in C. Here, plate voltage is applied at the center of the inductor, which is grounded by a bypass capacitor at that point. A single-section tuning capacitor is used, insulated entirely from d-c or a-c ground. An r-f peak of twice the d-c plate voltage appears across the tank. Many other variations are possible following the general principles used in each example.

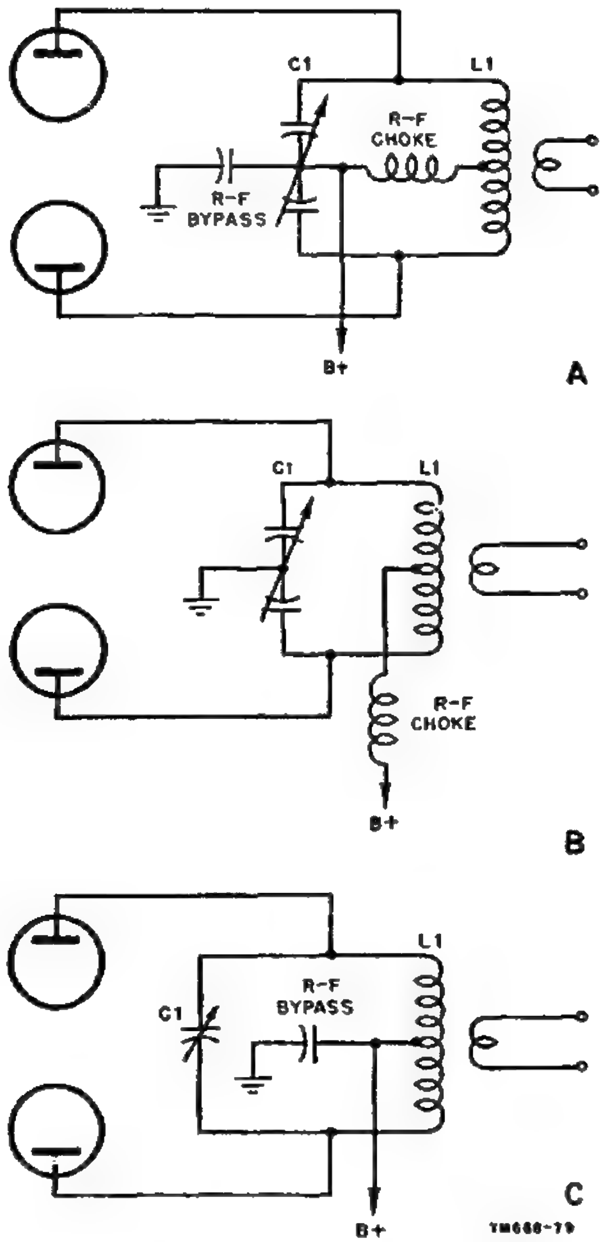


Figure 79. Push-pull plate tank circuits.

j. Antenna Matching Tank Circuits.

(1) In small, portable transmitters, it is common to find the antenna connected directly to the tank circuit of the transmitter with no intervening transmission line. Because the impedance of this kind of antenna can vary over a considerable range, it must be matched directly to the power amplifier by means of a tank circuit that can compensate for a wide range of impedance. Three basic circuits, adaptable to single-ended and push-pull tank circuits alike, which provide this variable impedance matching are shown in figure 80.

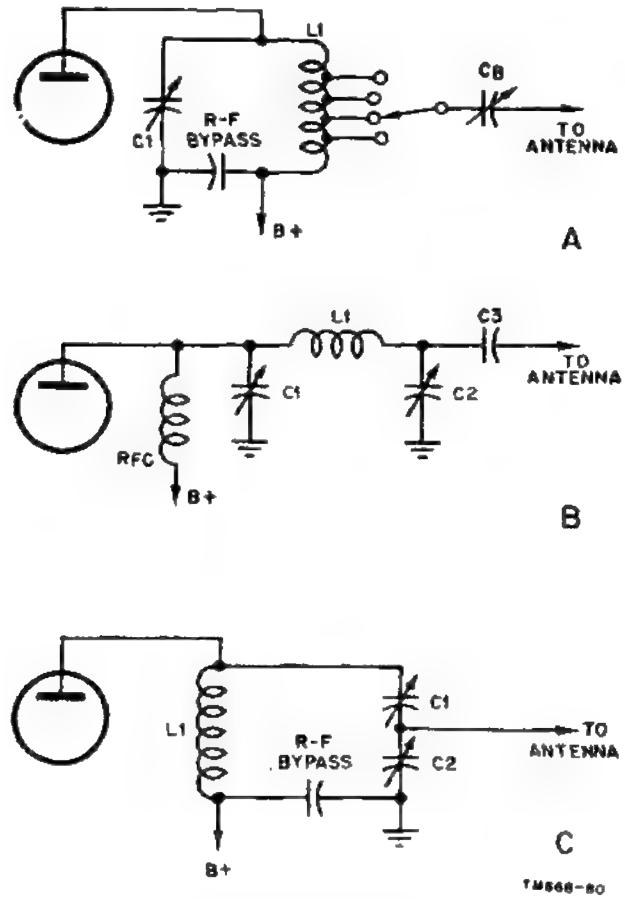


Figure 80. Antenna-matching tank circuits for power amplifiers.

(2) A series-fed, parallel-resonant, single-ended tank with grounded tuning-capacitor rotor and bypassed inductor is illustrated in A. Instead of coupling the antenna inductively to the tank, it is tapped directly to the coil through a blocking capacitor. Since the lower end of the tuned circuit is grounded effectively, the impedance to ground at that point must be zero. As the tap is moved up on the coil, the impedance rises until it reaches the ultimate value of the tank circuit impedance. In a

practical transmitter, the coil is tapped at intervals and a rotary switch is used to select the tap which gives the proper value of coupling. Because the d-c plate current in the amplifier stage increases with increased loading, the tap can be set at the point which gives the required d-c current in the stage when the tuned circuit is resonated. If the blocking capacitor, C_B, between the tap and the antenna is made variable, a further adjustment between separate taps can be obtained.

(3) One of the most frequently encountered variable matching networks for the output of an r-f amplifier is the circuit in B. The plate voltage fed through the r-f choke is prevented from reaching the antenna by blocking capacitor *C*3. The simple pi-network of *C*1, *L*1, and *C*2 is capable of matching a wide range of impedances, and operates as a voltage divider. The combination of *L*1 and *C*2 forms the divider circuit which develops higher or lower voltages at the output terminal. *C*1 then tunes the combination of *C*2 and *L*1 to resonance at the operating frequency. Depending on the relative values of *C*1 and *C*2, a voltage much lower than the a-c plate voltage can be developed. Consequently, this circuit can match an extremely wide range of impedances. In addition to matching purely resistive loads, the circuit also can compensate for a certain amount of reactance. This is important when using short antennas which introduce considerable capacitive reactance.

(4) A variation of the pi-network, in which one of the capacitors is not grounded, is shown in C. Capacitors *C*1 and *C*2 themselves form the impedance-matching voltage divider. The circuit cannot match as wide a range of impedances as the pi-network can, and it is further limited because the rotor of *C*1 must be carefully insulated. The response of these circuits to harmonics of the fundamental frequency is poor, which is a desirable feature. The pi-network does not discriminate against signals below operating frequencies. This makes it undesirable to use if the amplifier is driven directly by a frequency multiplier.

k. Parasitic Oscillation, Adjustment, and Neutralization.

(1) Many different types of input and output circuits for class C f-m power amplifiers have been described. At first glance, the large variety of choices available makes it seem difficult to understand why a particular circuit is chosen. Not all combinations of input and output circuits can be used together successfully, since some of them permit the amplifier stage to oscillate at frequencies that are relatively unrelated to the frequency to which it is tuned. These *parasitic oscillations* are distinct from the sort of oscillation that occurs in an amplifier which is improperly neutralized or one in which the input circuit is not shielded sufficiently from the output. They are undesirable because they cause the transmission of spurious signals, thus impairing the efficiency of the amplifier.

(2) The most noticeable features of parasitic oscillation in an amplifier are erratic tuning and the radiation of spurious frequencies. When an amplifier is operating properly, the d-c plate current dips sharply as the tank circuit is tuned through resonance. This plate current minimum also corresponds to maximum power output. If a tetrode is operating normally, the plate-current change may not be too great, but the screen-current dip will be significant. With parasitic oscillation, the plate current may not dip at all; the minimum may not correspond to maximum power output; or several dips may appear in the tuning range. Since the symptoms presented by a stage which is not properly neutral-

ized are somewhat similar, it is difficult to tell the two effects apart unless neutralization is checked first.

(3) All parasitics are attributable to the development of resonant circuits in connection with the tube elements in such a way as to permit enough feedback to sustain oscillation. They may occur at either high or low frequency. Parasitic oscillations occurring at much lower than operating frequencies usually are caused by the resonant condition of an r-f choke in the circuit, since the r-f chokes are the only inductors with sufficient inductance to resonate with various circuit capacitances at low frequencies. High-frequency parasitics can be traced to a much wider variety of causes. Among these are spurious high-frequency resonant conditions in tank-circuit inductances; resonant circuits built up in lead inductances and stray, or tube, capacitances, and resonant conditions built up in bypass and blocking capacitors. Moreover, the parasitic circuit need not involve the final amplifier alone. The driver stage is frequently an important part of the parasitic feedback circuit which permits oscillation.

(4) A recurrent type of high-frequency parasitic oscillation is caused by a form of tuned-plate, tuned-grid oscillator in a simple single-ended amplifier like that of figure 81. The parasitic path is shown in heavy lines. At relatively high frequencies, the tank-circuit inductance acts like an r-f choke, and the capacitors and the leads from them form the equivalent of parallel-resonant circuits. The shielding effect of the screen grid in a tetrode is not sufficient at extremely high frequencies. Therefore, energy can feed back to the grid circuit from the plate at high frequencies if both of the parasitic resonant circuits are almost the same in frequency. The difficulty can be cured by inserting a parallel inductance and resistance in the grid or plate lead. This detunes one of the parasitic circuits sufficiently to prevent oscillation. Another method is to insert a small resistance in series with circuit leads to introduce sufficient loss to stop oscillation. A third alternative is to incorporate a tuned parallel-resonant trap that actually inserts a very high impedance in the parasitic frequency path. In addition to the trap circuit, it is common to find small high-frequency capacitors connected from plate and control grid to cathode. These capacitors effectively bypass the harmonic path.

(5) Certain circuit combinations have been found to be troublesome. For example, r-f chokes rarely are used in both the grid and the plate circuit of a triode, since they cause a low-frequency tuned-plate, tuned-grid oscillation (B of fig. 81). For this reason, shunt-fed circuits are avoided whenever possible, since they encourage parasitic difficulties. In high-gain screen-grid amplifiers, the selection of the screen bypass capacitor becomes very important. The substitution of a different type when servicing a unit often leads to serious instability. Similarly, the choice of cathode or filament bypass capacitors is also a more critical matter than the circuit diagram tends to indicate. When replacing any of these components in a transmitter, use the *exact duplicate* of the discarded component, and *pay careful attention* to lead dress and parts placement.

(6) In push-pull circuits, there are many more possibilities for the development of parasitic oscillation. However, the oscillation usually can be traced to one of the standard types of oscillators operating in conjunction with the tube leads or r-f chokes. Remedies similar to those found in single-ended stages are used. An attempt always is made to space the spurious resonant circuits in such a way that none of them occur at the same frequency in the output and input circuits. As such,

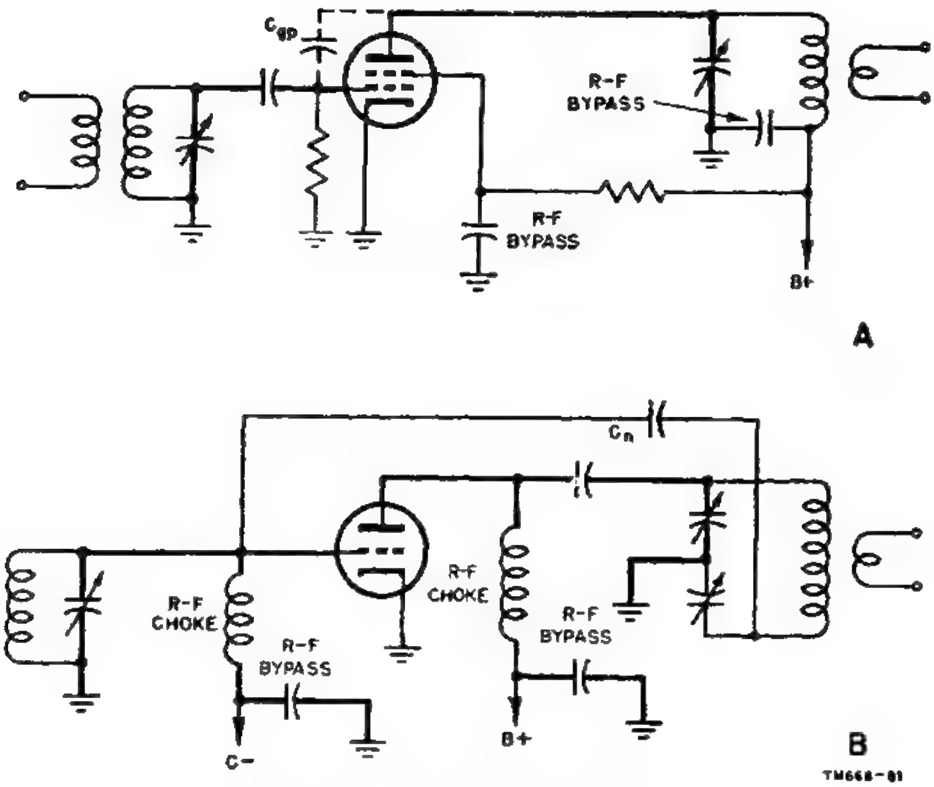


Figure 81. Parasitic oscillation circuits.

the circuit diagram of a power amplifier does not always tell the whole story of its operation. Many features are involved in the stability of the circuit that do not appear in the formal schematic, since they have to do with lead length, choice of wire size, placement of parts, and type of component.

(7) In f-m transmitters, improper neutralization can lead to a number of difficulties. The amount of frequency deviation can be changed by a final power amplifier on the verge of oscillation. In addition, such amplifiers are difficult to adjust for optimum power output and performance. Since the f-m signal is steady in amplitude, there should be no fluctuation of any of the d-c voltages or currents in the final amplifier when the transmitter is modulated. Any variation in rectified grid current indicates either overmodulation or instability in the driver stages. Fluctuation in d-c plate current points to parasitics or improper neutralization. Only a stable, properly tuned amplifier is capable of providing satisfactory performance.

Section II. AUTOMATIC-FREQUENCY CONTROL

42. Frequency Control

a. Description and Purpose. In military transmitters and receivers precise maintenance of assigned frequencies is imperative. Some transmitters must be capable of continuous frequency variation over an assigned band of frequencies and often are combined with receivers in the same unit. Since these receivers must be kept on exactly the same frequency as the transmitter, a control system must be used to compensate for any variation in frequency caused by vibration, temperature change, or humidity.

The control system is essentially an error-correcting system and usually is referred to as automatic-frequency control, or afc.

b. General Afc Systems. In an afc system, some of the output voltage is sampled and compared with a constant-frequency source, and any frequency difference that exists results in automatic correction of the frequency of the master oscillator. The functional arrangement of all afc systems is pictured in the block diagram of figure 82.

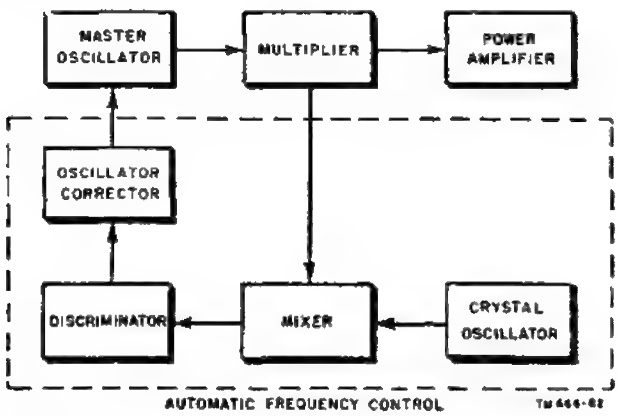


Figure 82. Functional block diagram of afc system.

(1) The classification of different types of afc depends on the means used to control the frequency of the oscillator. Oscillators which are frequency-modulated by a reactance or Miller effect respond to d-c voltages, and produce corresponding frequency changes. The correction voltage that is applied to the oscillator must be a d-c voltage directly proportional to the frequency difference between the output of the master oscillator and a standard crystal oscillator. The comparator, therefore, must produce a d-c voltage that corresponds to the difference frequency. Such a device is called a *discriminator.*

(2) Control of the frequency of an oscillator through the application of direct voltage to the modulator has certain disadvantages from the standpoint of precision. D-c amplifiers are unstable in respect to changes in tubes and line voltages. Moreover, if the control system fails, the frequency of the oscillator changes drastically because of the abrupt d-c voltage change at the modulator. From the standpoint of reliability this is undesirable. However, the d-c control systems are simply constructed and, despite their inherent disadvantages, they are widely used.

(3) To overcome the disadvantages of the d-c control system, a mechanical element can be inserted in the system to act as the oscillator frequency controller. This is usually a two-phase motor, whose rotor position depends directly on the phase relation of the voltages across its field windings. The motor has a shaft with a variable capacitor or inductor attached. The capacitor or inductor is connected in the circuit so that the reactive element varies the frequency of the master oscillator. This method of afc usually is applicable only to large fixed stations where weight and size are not important. For extremely high frequencies, where d-c control systems do not provide the required accuracy, the use of a motor-positioning system is frequently the only alternative. The motor positioning system has the advantage, in the event of failure of the control system, that the motor shaft does not turn. Therefore, the correction reactance produced does not change, and the frequency is not disturbed. Of course, if the failure lasts over a considerable period of time, the oscillator itself shifts frequency because of temperature changes and other causes.

(4) Motor control systems and d-c systems have varying speeds of response to errors of frequency. Because of the mechanical inertia of a motor, motor systems generally cannot respond quickly to changes in frequency such as the change caused by modulation of the carrier. However, it can respond much more accurately to slow variations. The d-c systems are pre-

vented from responding to modulation by use of circuits with long time constants.

(5) Receiver interlock systems, used in small portable units with single-dial control, are generally of the d-c type. Interlock systems use the same master oscillator for the receiver and the transmitter. Mixer circuits using crystal-controlled oscillators produce the required frequencies for the local oscillator of the superheterodyne receivers and for the actual master control of the transmitters. This system can be further refined by using the receiver to keep the transmitter locked to the frequency of another station.

43. Discriminator

a. General.

(1) The *discriminator* is a device for producing a d-c voltage which is proportional to the frequency of an input signal. The polarity of the voltage produced depends on whether the frequency is higher or lower than the frequency to which the discriminator is tuned. The response curve of a tuned resonant circuit (fig. 83) shows that the sides of the curve approach straight lines as the Q of the coil is increased. If a signal of variable frequency is coupled to a tuned circuit, the voltage produced across it will depend on the relation of the frequency of the coupled voltage to that of the tuned circuit. The voltage also depends on the Q, since Q defines the sharpness of resonance so that frequencies farther away from resonance produce less voltage across the circuit.

(2) If the a-c voltage that exists across a resonant circuit is rectified by a diode, a d-c voltage proportional to the amplitude of the a-c voltage is produced. If the amplitude of the a-c voltage varies with the displacement of the applied frequency from the actual resonant frequency of the tuned circuit, the d-c voltage will increase or decrease.

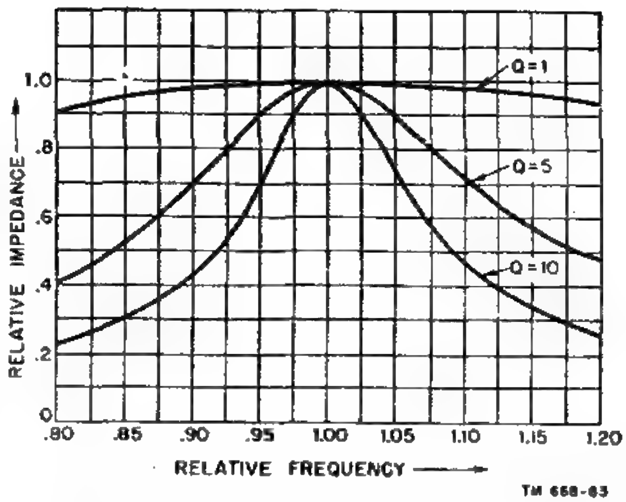


Figure 83. Response curves of tuned resonant circuits.

b. Double-Tuned Discriminator.

(1) Figure 84 shows a double-tuned discriminator consisting of tuned circuits *T*1, *T*2, and *T*3, diode rectifiers *D*1 and *D*2, and the filter networks, *R*1*C*1 and *R*2*C*2. The secondaries, *T*2 and *T*3, are tuned to resonate at different frequencies; one is tuned above the carrier frequency and the other an equal distance below the carrier frequency. This provides equal voltages at the center frequency, as shown in the response curve of figure 85. When an r-f voltage that is constant in amplitude and varying in frequency is applied to *T*1, the voltages induced in *T*2 and *T*3 will be 180° out of phase, and alternate voltage polarities will appear at the plates of *D*1 and *D*2. These induced voltages increase and decrease in amplitude with the changing frequency. For example, assume that *T*2 is tuned to a frequency higher than the center frequency. As the induced voltage approaches the resonant frequency of *T*2, its amplitude increases in a positive direction. If *T*3 is tuned to a lower frequency than the center frequency, as the induced voltage approaches the resonant frequency of *T*3 its amplitude increases in a negative direction. When the induced voltage goes positive at the plate of *D*1, current flows in the circuit *D*1, *T*2,

and $R1$ and a voltage proportionate to the change of frequency appears across $R1$. As the induced voltage goes positive at the plate of $D2$, current flows in circuits $D2$, $T3$, and $R2$, and a voltage proportionate to the change in frequency appears across $R2$. Capacitors $C1$ and $C2$ across the rectifiers filter out any a-c variations, and permit a d-c voltage to be built up across the load resistors.

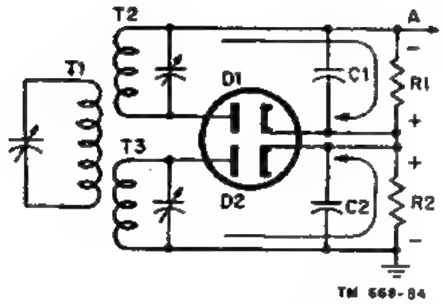


Figure 84. Double-tuned discriminator circuit.

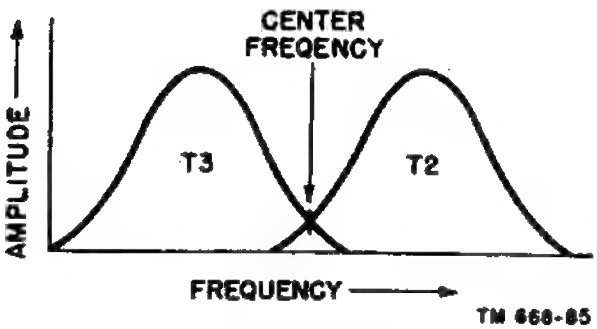


Figure 85. Response curve of double-tuned discriminator.

(2) With the connection of the rectifier as shown, the voltages across the individual load resistors oppose one another because the cathodes are both at the same potential. Therefore, the total voltage between the top of $R1$ and ground depends on the relative value of the voltages across $R1$ and $R2$. Since the voltage across the individual load resistors depends only on the frequency of the applied signal, when the frequency is higher than the center frequency, the voltage developed by the diode connected to the circuit tuned above the center frequency is higher than that developed across the other tuned circuit. Similarly, if the applied frequency is lower than the center frequency, the diode connected to the low-frequency tuned circuit produces the larger voltage across its load resistor. If the frequency of the applied signal is exactly at the center frequency, the voltage across the load resistors is equal, and the total output voltage is zero.

(3) When the applied frequency is higher than the center frequency, more voltage is developed across $T2$, a greater d-c voltage appears across $R1$, and A becomes more negative. With the carrier at the center frequency, the voltage at A, in respect to ground, becomes zero. When the frequency swings lower than the center the voltage developed across $T3$ is greater than that produced across $T2$, a greater d-c volt-

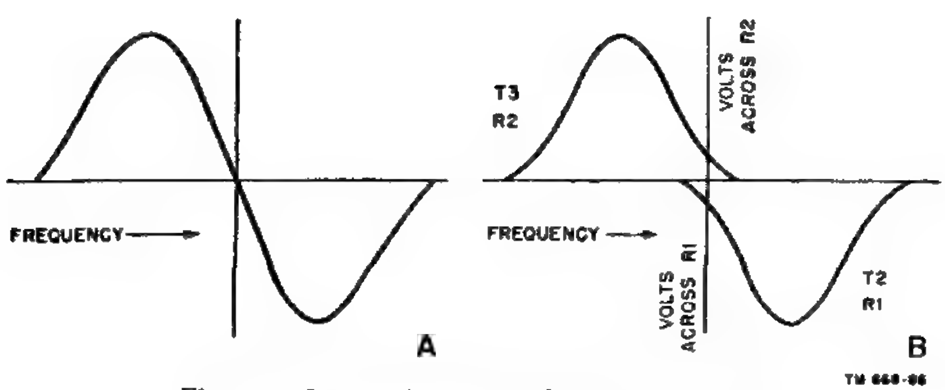


Figure 86. Output voltage of double-tuned discriminator.

age appears across $R2$, and point A becomes positive in respect to ground. Therefore, as the applied signal swings from below to above the center frequency, the voltage at A goes from positive to zero to negative. This results in the curve of output voltage versus frequency shown in A of figure 86. The voltage across the individual load resistors in respect to frequency is shown in B.

c. Use of Double-Tuned Discriminator.

(1) The potential developed at the output of a double-tuned discriminator can be used as a frequency-correction voltage by applying it to the grid of a reactance-modulator tube. When the discriminator voltage changes polarity, the transconductance of the tube is increased or decreased and the frequency of the oscillator shifts. For example, assume that the reactance modulator is connected to inject capacitance into the oscillator circuit. When the transconductance is reduced, less capacitance is injected and the oscillator frequency increases a small amount. As the frequency applied to the discriminator changes and the voltage output becomes positive, the transconductance of the reactance modulator is increased. This injects more capacitance across the oscillator tank circuit and the frequency of oscillation is lowered.

(2) If the discriminator is tuned so that an increase in frequency of the oscillator produces a positive voltage, the discriminator voltage applied to the reactance-modulator tube tends to return the oscillator to the center frequency. Similarly, a decrease in frequency will cause the oscillator to return to its normal frequency. Since all of the operations take place at the oscillator frequency, and since the frequency at which the system becomes stable is the center frequency of the discriminator characteristic, this is a crude control system. The inductance and capacitance in resonance are no more likely to be stable than the inductance and capacitance of the oscillator tank circuit itself. However, this is the only method of afc that can be used in some ultrahigh-frequency systems. Because the over-all accuracy is totally dependent on the center of the discriminator frequency characteristic, the discriminator must be more stable than the oscillator for accurate frequency control.

(3) A high-accuracy afc system should keep the center frequency of the master oscillator as stable as a crystal oscillator. This can be accomplished by comparing the frequency of the applied signal with the frequency of a crystal. The frequency difference between the crystal and the master oscillator, as produced in a mixer, depends on the absolute values of both frequencies. Therefore, if the crystal oscillator is assumed to be absolutely stable, the difference frequency depends only on that of the transmitter frequency. This difference frequency is applied to a discriminator operating at a much lower frequency than the oscillator and the transmitter. If the frequency at which the discriminator operates is made low enough, the result of a variation in its tuned circuits is only a few kilocycles. Therefore, if the crystal frequency, or any multiple of it, is mixed with the master-oscillator ouput to produce a low difference frequency, the low-frequency discriminator tuned circuits will cause only a small error, whereas the over-all variation of difference frequency caused by drift in the oscillator will be much larger. The departure of the transmitter frequency from the difference frequency produced by the mixer determines the correction voltage. The output of the discriminator feeds the reactance modulator tube, which brings the master oscillator back to the center frequency. The over-all stability is nearly that of the crystal oscillator itself, differing only by the de-

parture of the discriminator itself from the low-center frequency. This is a departure of approximately a few hundred cycles per second.

(4) The frequency of the master oscillator shifts with the applied audio signal, but the discriminator is so constructed that it responds only to changes in frequency that are much slower than the audio variations. The voltage at the output of the discriminator varies at a rate equal to the rate of change of frequency of the master oscillator—that is, at the audio rate. It also varies at a much slower rate because of oscillator drift. By placing a low-pass filter after the discriminator, only the slow drift variations can reach the reactance tube. This filter cuts off all voltage changes with a rate that is equal to or higher than the audio frequencies used.

d. Phase Discriminator.

(1) Figure 87 shows the schematic diagram of another type of discriminator circuit used in afc circuits. This is known as the phase discriminator. The voltage output of the discriminator depends on the phase relations in the circuit. Transformer *T*1 couples the input from the crystal oscillator to the diode plates producing equal and opposite voltage *E*2 and *E*3 on these plates. The phase relationship of these two voltages is shown in A of figure 88. The crystal oscillator and transmitter frequencies are reduced so that these voltages are generally of a frequency just above the audio range. Since the input to *T*1 is crystal-controlled, it can be considered as stable. Therefore, the frequency and relative phase of *E*2 and *E*3 never change. The input from the master oscillator of the transmitter is injected into the primary of *T*2. This produces a voltage, *E*1, across its secondary, which is exactly 90° out of phase with *E*2 and *E*3 when the frequency of the input from the crystal is the same as the frequency from the master oscillator. The upper diode receives a voltage equal to the sum of *E*1 and E2; the lower diode receives voltage equal to the sum of *E*1 and *E*3. The diodes rectify the signals and d-c voltages appear across the load resistors. When the signal voltages are equal, the output voltages across the resistors are equal, and the total voltage across both resistors in respect to ground is canceled out.

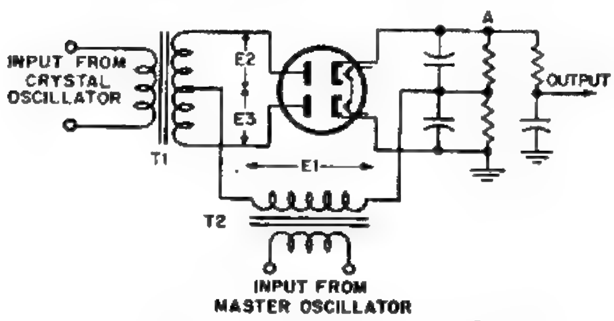


Figure 87. Basic phase discriminator.

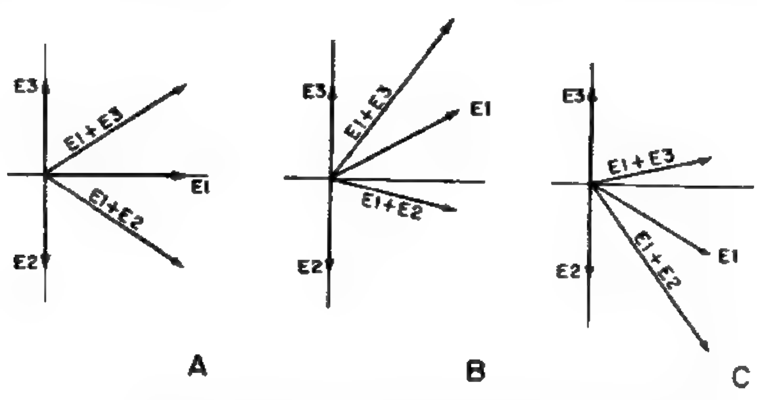


Figure 88. Vector relations in phase discriminator.

(2) If the frequency of the master oscillator increases, the vector relationships change as shown in B. The increased frequency is equivalent to a phase shift of $E1$ in respect to $E2$ and $E3$. The resultant vectors, $E1$ plus $E3$ and $E1$ plus $E2$, therefore change in length, as shown, and the ratio of the voltage across the diodes changes. The voltages across the load resistors (fig. 87) are now unequal and a voltage that can be fed back into the modulator to correct the drift of the master oscillator is produced at point A.

(3) Similarly, if the frequency of the master oscillator decreases, the phase of $E1$ changes, as shown in C of figure 88 and a voltage of the opposite polarity is produced across the diodes. The voltage across the lower diode is now less; therefore, a less positive voltage is produced at point A (fig. 87). The result is a change in the polarity of the over-all voltage to that produced when the frequency of the master oscillator shifts upward in frequency. The correction voltage applied to the proper circuits tends to return the system to a condition where $E1$ is 90° out of phase in respect to $E2$ and $E3$. Consequently, the frequency of the master oscillator is dependent on that of the crystal oscillator. Since the discriminator depends only on phase relationships and not on the absolute frequency of the secondaries of the transformers, the tuning of the discriminator has less effect on the stability of the system. However, divider circuits are needed to reduce the transmitter and crystal oscillator frequencies and they are considerably more complex than the simple mixer system.

e. Modified Phase Discriminator.

(1) The complex divider circuits necessary have led to a modification in the basic phase discriminator. Essentially, this modification consists in obtaining the voltage that corresponds to that produced by $T2$ in figure 87 from the same circuit to which the split secondary of $T1$ is connected. This modification is shown in figure 89. A tuned circuit is used for the split transformer, the signal voltages across the two halves of the secondary are still $E2$ and $E3$, and the diode load circuits and their operation are the same as before. The voltage present across the primary winding is also present across L and the voltage, $E1$, is obtained from across this inductor. Therefore, the secondary system receives its voltages in two ways—by inductive coupling, and through coupling capacitor C. The voltage is the same as the voltage, $E1$, across the primary and is 180° out of phase with the total voltage, $E2$ plus $E3$, induced across the secondary circuit. The voltage drops across each half of the split secondary are, in turn, 90° out of phase with the applied voltage. However, they are 180° out of phase with each other, since the circuit is a tapped transformer, and therefore, the voltages $E2$ and $E3$ are developed 90° out of phase with the primary voltage.

(2) When the frequency of the carrier departs from the center of the tuned-circuit resonance curve, the phase relationships across the tuned circuit change, and the over-all d-c output changes. Consequently, the d-c output of the discriminator varies with the applied frequency. However, operation is dependent on the resonant frequency of the discriminator secondary, although this discriminator is much easier to tune and adjust than the double-tuned discriminator, because only one resonant circuit is used.

f. Pulse Discriminator.

(1) Another type of discriminator that can be used in afc systems—the *pulse discriminator* — uses pulses rather than continuous sine-wave signals. The pulse discriminator distinguishes between two pulse signals of different repetition rates, and produces a volt-

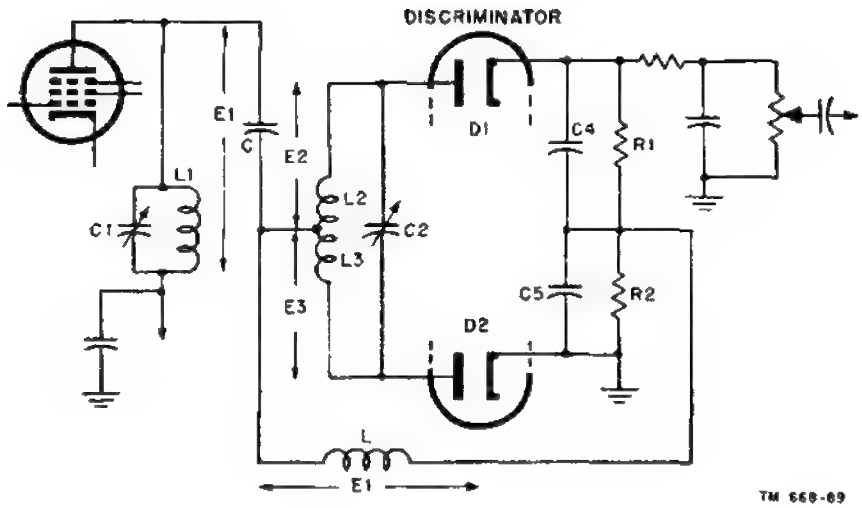


Figure 89. Modified phase discriminator.

age pulse proportional to their difference. It is necessary to convert the frequency of the master oscillator to a series of pulses with a repetition rate proportional to frequency. The repetition frequencies are compared in the pulse discriminator and an output is produced if the repetition rates differ. This output is converted to a d-c voltage and applied to the modulator, thereby shifting the frequency and bringing the two repetition rates back into synchronism.

(2) The accuracy of the pulse system is much greater, in terms of timing, than that of a system using sine waves. A pulse begins at a precise instant, whereas a sine wave does not, because the rate of voltage change at the beginning of a pulse is much greater than the rate of voltage change at the beginning of a sine wave. A pulse discriminator responds to smaller changes in the repetition rate than an equivalent phase discriminator, which responds to changes in the frequency of the input signals. This system of stabilization has the advantage of requiring no frequency-divider circuits. Moreover, no tuned circuits of any kind are used in the actual stabilization loop. This means that there need be no tuning or operating adjustment of the stabilization circuits.

(3) The over-all block diagram of the stabilization circuit is shown in figure 90. Some of the f-m signal from one of the multiplier stages in the transmitter is tapped off and fed into a buffer amplifier. This amplifier applies the signal to the input grids of two mixer tubes. A crystal oscillator, operating at the same frequency as the input signal, is applied also to the mixer grids. However, the signal from the oscillator is passed through two R-C networks, each of which shifts the phase by 45°. To accomplish this, the value of the resistance in each R-C network is made equal to the reactance of the capacitor at the crystal frequency. The upper network shifts the phase forward by 45° and the lower retards it by the same amount. Therefore, the dual output of the networks differs in phase by 90°.

(4) The mixer plate circuits select the frequency which is the difference between the signal from the buffer amplifier and the signal from the oscillator. When no modulation is applied to the transmitter and the frequency of the master oscillator is stable, the frequency from the buffer oscillator and

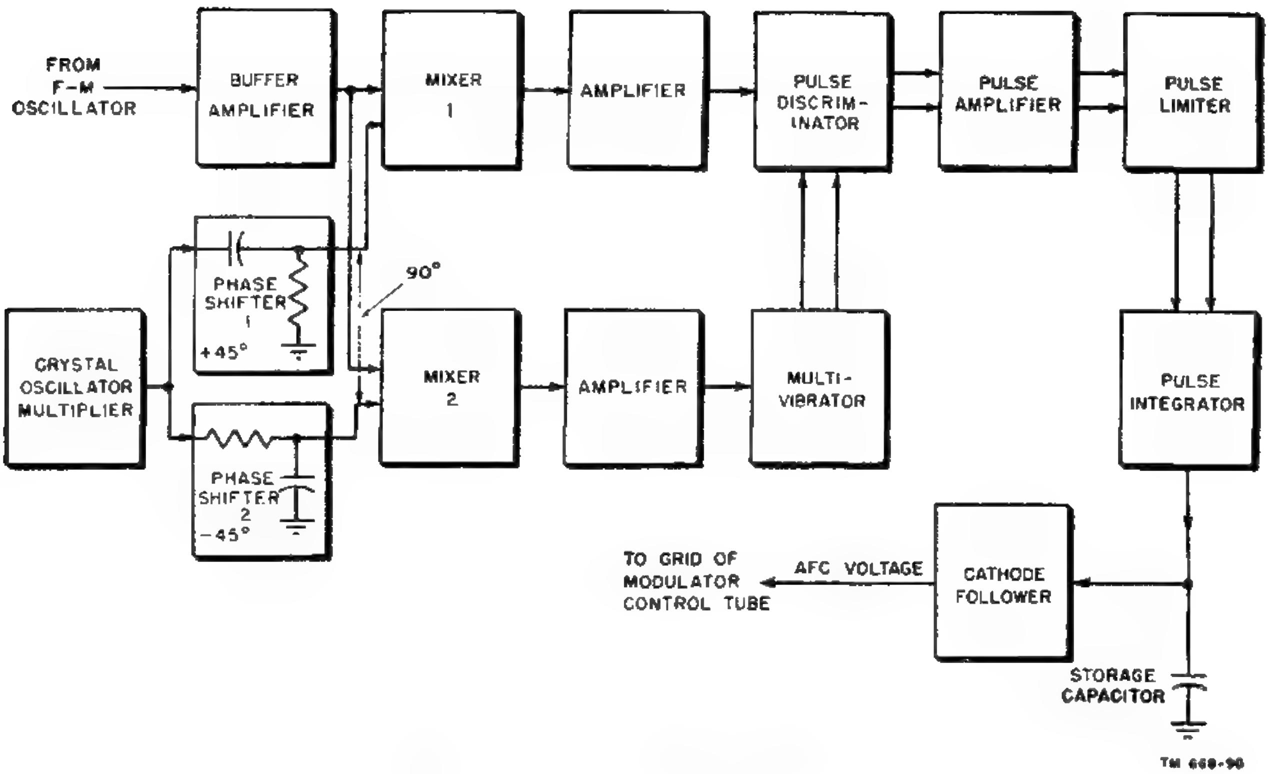


Figure 90. Afc system using pulse control.

that from the crystal oscillator are the same, and there is no output from the mixer because the difference frequency between the applied signals is zero and and the mixer is set to accept only the difference frequency. When the difference frequency is present, it is always very small compared to the frequency of either of the applied signals.

(5) When modulation is applied, the frequency of the master oscillator increases or decreases. Therefore, the input signal to the grids of the mixers will increase or decrease by a like amount. This signal mixes with that from the crystal oscillator and the difference frequency is equal to the instantaneous deviation of the mixer. When the modulated oscillator is on frequency, the upper and lower halves of each cycle of difference frequency contain the same average amount of area over a period of time. If the oscillator drifts upward, the output of the mixers is a wave in which average power in the upper half of the wave is larger than in the lower half. The control circuit is used to keep the average power in the upper and lower halves of the cycle equal by applying a suitable correction voltage to the master oscillator.

(6) Because of the phase-shifting networks, the output from the first mixer lags that of the second mixer by 90° when the frequency of the master oscillator is higher than that of the crystal. When the frequency of the oscillator is lower than that of the crystal, the output from the first mixer leads that of the second. At all times, the output of both mixers is 90° out of phase because of the constant difference in phase at the output of the crystal phase-shifting networks. During 1 cycle of modulation, the frequency of the oscillator first increases and then decreases. Therefore, the output of the mixers varies from minus 90° to plus 90°. This happens on each half-cycle.

That is, the phase lags on the high half-cycle of modulation and leads on the low half-cycle (fig. 91); f_o is the frequency of the oscillator and f_c is the center frequency.

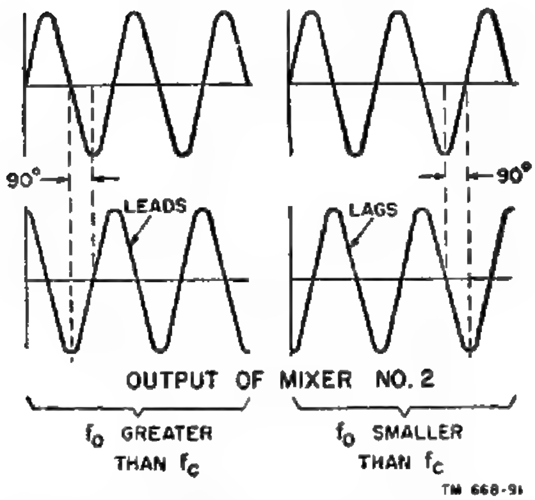


Figure 91. Phase relations in balanced modulators used in pulse control systems.

(7) The output from each mixer is passed through a separate amplifier. One amplifier is biased so that a few volts of negative signal drive it completely into cut-off. Therefore, the plate voltage increases to the value of the supply voltage and stays that way until the voltage on the grid allows the tube to conduct. When the grid swings positive, it draws current, and drives the tube to saturation. The plate current cannot increase and the plate voltage stays constant until the voltage on the grid goes negative. The result is the production of a square wave. The output from the other amplifier is used to trigger a multivibrator circuit that also develops a square wave output. This circuit acts as a switch and turns the d-c supply voltage off and on in response to the input signal. Each positive half-cycle develops a corresponding pulse in the output.

(8) The multivibrator circuit produces two square waves that are 180° out of phase with each other. The out-of-phase square waves are passed through R-C differentiating networks, as shown in figure 92. These differentiating networks are chosen so that the capacitance is small compared to the resistance. Therefore, the reactance of the capacitor is large at low frequencies and only high frequencies are passed. The square wave can be considered as composed of a number of sine waves of various frequencies and phase relationships. The highest of these occur at the leading and trailing edges of the wave. Therefore, the R-C networks produce pulses that are sharp and narrow.

(9) The pulses produced by the leading and trailing edges of the square waves are applied to the plates of a dual-diode pulse discriminator. The square waves from the first mixer and amplifier are applied to the junction of the two resistors. The cathode is biased negative to an amount equal to the peak value of the out-of-phase square waves. Therefore, neither the out-of-phase pulses nor the clipped signal from the first mixer can make the diodes draw current. However, when they are both present at once and in phase they are effectively in series at the diode plates.

(10) In figure 92, diode *D*1 is conducting because the voltages are adding in the wrong direction on *D*2. The output signal from *D*1 of the pulse discriminator is in the form of short positive pulses. Because of the bias in the output, only signal amplitudes larger than the bias can appear. Therefore, the actual output is not the square wave plus pulses, but the pulses alone. The discriminator produces positive pulses from *D*1 when the frequency of the master oscillator is above that of the crystal oscillator and negative pulses from *D*2 when it is below. The repetition rate of the pulses depends on the amount by which the two oscillators differ in frequency.

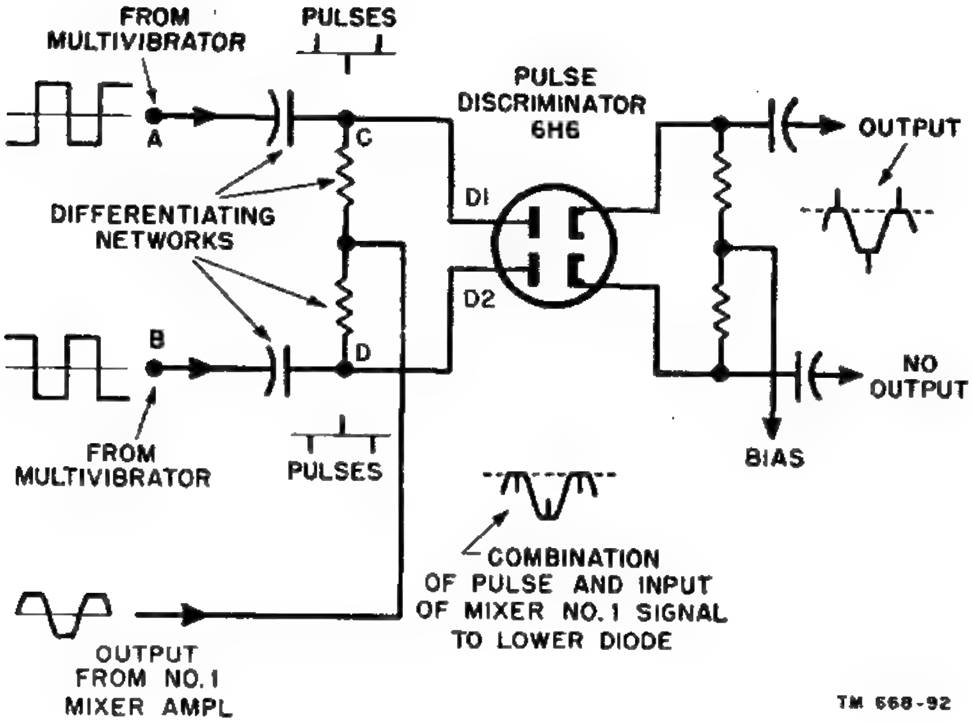


Figure 92. Pulse discriminator.

(11) The pulses are made uniform in amplitude by passing them through an amplifier and a pulse limiter, as shown in figure 90. They then are applied to a capacitor in the pulse integrator and each pulse charges the capacitor a small amount. When the frequency of the master oscillator swings high on modulation, more positive pulses are produced, and the capacitor charges to a positive voltage. On the downward swing of the carrier the reverse is true, and the capacitor charges negatively. If the upper and lower swings are equal about the constant value of the crystal oscillator, the net charge across the capacitor is zero. However, when the oscillator drifts higher or lower, there is an increase in the number of negative or positive pulses. This alters the net charge on the capacitor and is equivalent to a changing voltage across it. This changing voltage is applied to the grid of the modulator tube to correct any drift in the oscillator itself.

(12) This system needs no tuned circuits, and stabilizes on alternate halves of the modulation swing. The response time of the system to changes in center frequency can be as low as the period of 1 cycle at the lowest audio frequency. Long-time variations also are compensated for, since these depend only on the number of pulses and not on changes in supply voltage or tubes. The over-all circuit can be used only in large fixed transmitters where the number of component parts is not a problem.

44. Motor Control Systems

a. Description. The motor control system uses a two-phase motor attached to a variable capacitor and connected across the master oscillator tank. The frequency-control circuit derives two voltages 90° out of phase to turn the motor if the frequency of the master oscillator does not coincide with the frequency of the crystal oscillator.

b. Circuit. The frequency-control section of a motor-control system consists of three main parts—the crystal oscillator, the frequency di-

vider, and the motor-control section. Some of the signal from the master oscillator is tapped off and fed to a chain of frequency dividers as shown in the block diagram of figure 93. The low-frequency divided signal is fed to two balanced-modulator circuits. The output of the crystal oscillator section is divided, so that the frequency is the same as that of the divided signal from the master oscillator. It then is split up, shifted in phase by 45° in each half, and applied to the balanced modulators. The output of the balanced modulators is fed to a four-winding two-phase motor. When the phase in all windings is the same, there is no rotation of the motor. If the master oscillator drifts, the phase in two of the windings differs from that in the other pair, and the motor rotates to correct the unbalance.

c. Balanced Modulator Motor Control System. The output of the crystal-oscillator divider system is split up by a phase-shifting network and applied to the grids of a balanced modulator. The input from the master-oscillator divider system also is applied to the modulator grids. When the two frequencies are equal, no output appears in the plate circuits of the modulator, but any difference between the frequencies of the input signals will cause an output to appear. This output is an a-c voltage with a frequency equal to the difference in frequency between the two input signals. Each balanced modulator output is connected to two windings of a four-winding, two-phase motor. When the frequencies of the crystal oscillator and the master oscillator in the modulators coincide, the difference frequency in the output is zero, and no voltage is applied to the motor. When a difference frequency is present, the output of one balanced modulator is 90° out of phase with the other. This 90° out-of-phase pair of voltages applied to the motor windings causes a rotating magnetic field to be set up. Therefore, the armature of the motor turns with speed of rotation equal to the speed of rotation of the magnetic field. This, is turn, depends on the frequency of the difference of the signals applied to the modulators. The motor, in turning, moves the capacitor connected across the oscil-

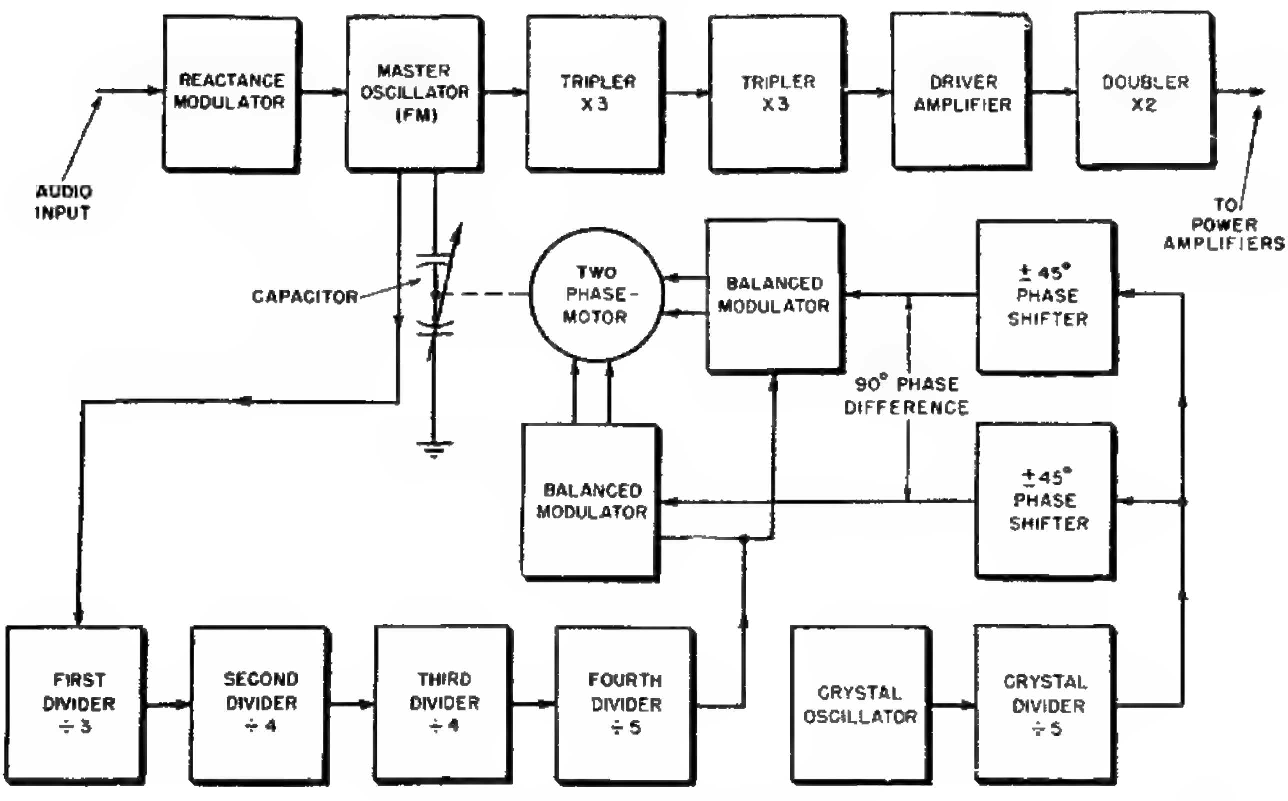


Figure 93. A motor-control afc system.

lator tank, so that the frequency of the master oscillator also changes. As the frequency changes, the difference between the master oscillator and the crystal frequency becomes less and less and the motor runs more slowly until it stops when the two frequencies coincide. Therefore, a small displacement of the oscillator frequency causes the motor to turn a very small amount and correct the drift.

45. Frequency-Divider Circuits

a. General. Frequency-divider circuits reduce the frequency of a signal by an integral multiple of the fundamental and also can divide and multiply by fractional quantities. They also reduce any deviation that is present on the signal because of modulation. There are two general types of frequency dividers, those that produce an output whether an input signal is present or not, and those which produce output only when the input signal is applied. The free-running dividers are all some variety of oscillator synchronized with a higher frequency. The other type of divider depends on the properties of special circuits and not only can produce division and multiplication, but also can divide and multiply by fractional quantities.

b. Multivibrator.

(1) One of the simplest oscillators that can be used as a frequency divider is the synchronized multivibrator. There are many varieties of multivibrator circuits, but essentially they are all modifications of a two-stage resistance-coupled amplifier circuit with the output fed back to the input circuit. When the grid voltage of a vacuum tube is made more positive the plate voltage decreases. This decrease in plate voltage is coupled into the grid of one tube, causing a decrease in grid voltage. This results in an increase in plate voltage, which is applied to the grid of a second tube, and the cycle reverses. The circuit is shown in figure 94. The variations possible consist in using direct coupling, cathode coupling, or mixed types of coupling between the two tubes.

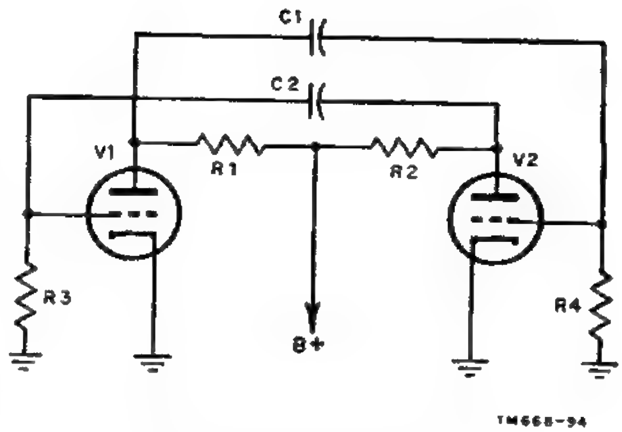


Figure 94. Simple multivibrator circuit.

(2) A small amount of voltage applied to the grid circuit can be used to trigger oscillation. Any voltage that is an integral multiple of the natural frequency of the oscillator provides this triggering action. The frequency can be much higher than the actual frequency of operation of the oscillator. The output from one multivibrator controlled in this manner can be ten times less in frequency than the controlling voltage. The output of this multivibrator can be connected to another multivibrator that also divides by a like amount, providing division by one hundred. In this way, the high frequency of the crystal oscillator and the master oscillator in an f-m system can be reduced to a frequency in the audio range. This can be applied to the phase or pulse discriminators for frequency control, as described previously.

(3) If the synchronizing voltage is not applied to the multivibrator, the oscillations do not stop, but run freely; hence, the name, *free running.* This is a distinct disadvantage in a divider circuit because if synchronism is lost, frequency control also is lost. For this reason a modification of the multivibrator known as the *one-shot* multivibrator or *trigger circuit* often is used (fig. 95). The circuit is essentially the same as the multivibrator, but

capacitors $C1$ and $C2$ of figure 94 are replaced by resistors $R5$ and $R6$. It sometimes is called a direct-coupled multivibrator. A small change in grid voltage of $V1$ increases the plate current. This increases the voltage drop across $R1$ and makes the grid of $V1_a$ more negative, decreasing the plate current through $R2$. As the grid of $V1$ becomes more positive, there is an abrupt increase of the plate current of the first tube and the plate current of $V1_a$ is cut off. Another pulse applied to the grid of tube 2 upsets this condition and causes another reversal, with maximum current in $V1_a$. Therefore, this circuit depends entirely on the input pulse, and does not operate (there is no output) when no pulse is present.

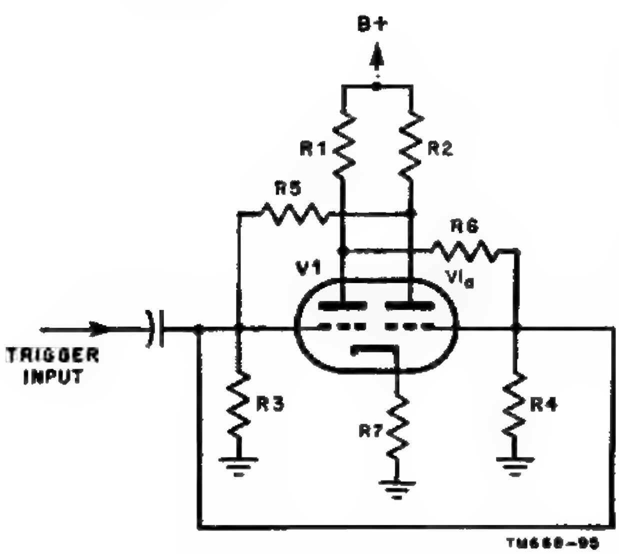


Figure 95. Trigger circuit.

(4) If several of the circuits described above are connected in sequence, the application of a pulse to the input of the first tube which was not previously conducting, conducts, and vice versa. The next stage is made to trigger only on the output of one conducting tube, and, since its plate voltage is low when conducting, the second stage does not trigger on the first pulse. A second pulse, applied to the first tube, triggers the second tube and causes it to conduct. A third pulse triggers the next tube in line and so on. If ten tubes are used, the tenth tube triggers on the tenth pulse. If this tube connects to a similar group of circuits, the last tube in the second line will trigger on every hundredth input pulse. This provides an accurate frequency divider that can be extended to very high orders of division. It has the advantage that there is no output unless an input signal pulse is present. On the other hand, a very large number of tubes becomes necessary if high division is needed.

c. Synchronized Oscillator. A vacuum-tube oscillator tends to synchronize with an injected voltage of about the same frequency. Also, if the frequency of the oscillator and the injected voltage are in approximate harmonic relationship, the oscillator synchronizes with the harmonic. For example, if a signal of 1 mc is injected into an oscillator operating at about 99 kc, the oscillator begins to oscillate at 100 kc, which gives a frequency division of ten. As the frequency stability of the oscillator is reduced, synchronization can be obtained over a wider and wider range. Synchronization can be improved if afc is applied to the oscillator. This is accomplished by applying the output of the oscillator and the voltage to be divided to a phase discriminator. The rectified output of the discriminator actuates a reactance modulator, which in turn changes the frequency of the oscillator. If a harmonic generator is inserted between the oscillator and the detector, the oscillator can be held in synchronism with its own harmonic, and therefore can act as an accurate divider. However, the circuit continues to function even with no synchronizing signal present. The oscillator therefore is free-running (fig. 96).

d. Regenerative Modulator. When two different frequencies are applied to a mixer circuit, the output contains all possible sum and difference frequencies and their harmonics. If one of the multiples of the difference is selected, amplified, and inserted back into the mixer circuit, it reinforces the output at that frequency. This is known as *regenerative modulation* (fig. 97). This device uses an ordinary

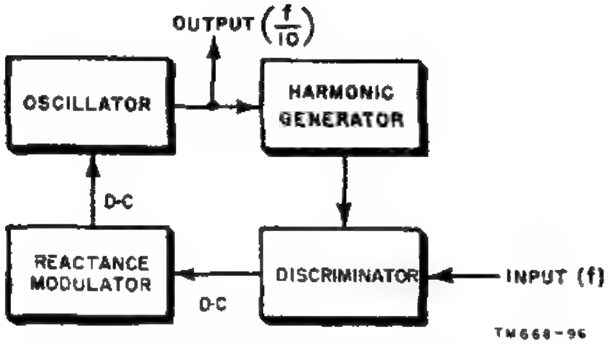


Figure 96. Synchronized-oscillator frequency divider.

mixer tube and a suitable selective amplifier. For example, the input to the mixer grid produces a distorted wave, which contains the tenth submultiple. The output is taken from the amplifier and fed back into the mixer, where it increases the amplitude of that submultiple and tends to suppress the others. The result is a divider that cannot operate unless an input signal is present. Fractional division can be obtained if the diagram of figure 98 is used. Here, the output of the amplifier is fed into an odd-harmonic generator, and an odd fractional value of the input signal is reinforced.

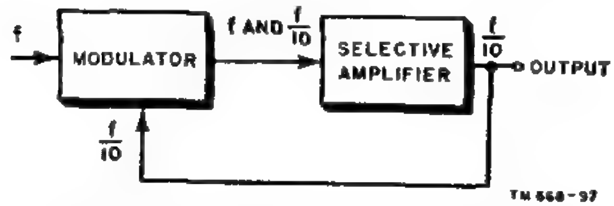


Figure 97. Regenerative-modulator frequency divider.

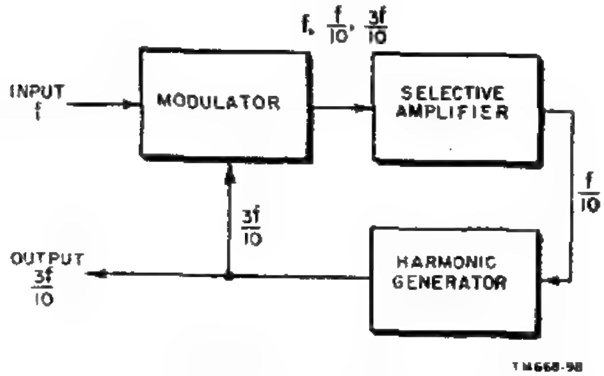


Figure 98. Regenerative-modulator fractional divider.

Section III. COMPLETE TRANSMITTERS

46. Indirect F-M Transmitter

a. The preceding sections of this chapter have discussed the operation of various circuits used in f-m transmitters. An over-all circuit diagram of these circuits combined in an actual piece of equipment would be large and complicated, especially if the transmitter has a large number of stages. An over-all schematic for a typical f-m transmitter, using indirect methods for producing the production modulation, is shown in figure 99. Although many things are shown at the same time, it is easy to follow the entire plan if the individual circuits that compose it are understood. For greater clarity, the power-supply wiring as well as the control circuits has been eliminated from this diagram. Normally, when printing the complete schematic, filament wiring, connections for plate voltage, and control circuits are all shown.

b. The indirect f-m transmitter has a Pierce crystal oscillator, V1, which is similar to the ultraudion oscillator. The crystal acts as a tuned parallel resonant circuit connected between the grid and the plate through a blocking capacitor, *C*28. Grid-leak bias is provided by the combination of resistor *R*1 and capacitor *C*1. Plate voltage is supplied through *R*2, and the output voltage of the oscillator is developed across it. The coupling capacitor, *C*2, couples the output voltage to the grid of the following stage.

c. The Link phase modulator is used in this stage, and the voltage drive for *V*2 is developed across grid resistor *R*3. Audio voltage is introduced through the correction network, *R*6 and *C*3, which produces the necessary frequency response required for the generation of true f-m. The high cathode bias that is used with this modulator is provided by *R*4, its value being great enough to operate the tube in a region of low transconductance. The frequency-modulated output is developed at the plate of *V*2.

d. The resistance-coupled a-f voltage amplifiers of the unit are formed by the circuits of *V*3 and *V*4. Included in the amplifier are a pre-emphasis circuit, a gain control, and decoupling circuits. The voltage developed by the microphone is isolated from the grid of the first audio tube, *V*4, by the microphone input transformer. At the same time this transformer provides a certain amount of voltage gain. Resistor *R*22 serves to stabilize the impedance of the secondary, so that the pre-emphasis network formed by *R*21 and *L*8 can function properly. Screen voltage and bypassing are provided by *R*19 and *C*27 respectively, and *R*20 is the conventional plate load resistor. Capacitor *C*24 in conjunction with resistor *R*15 acts as a decoupling network, to prevent feedback from the following stages from returning to the grid of *V*4 through the common power supply impedance. The same function is performed by *R*14 and *C*23 for the following stage. In this stage, *R*17 and *C*25 are the conventional cathode bypass and bias circuit, and *R*16 serves as the plate load resistance. Control of the voltage, fed from *V*3 through coupling capacitor *C*26 to the modulator, is achieved by variable resistor *R*18 in the grid circuit of *V*3. The output of the amplifier stages is supplied to the audio correction network, *R*6 and *C*3, through coupling capacitor *C*4.

e. From the modulator stage, *V*2, the frequency-modulated signal passes through capacitor *C*5 and builds up a voltage across *R*7 in the grid return of *V*5. This stage is a class A buffer amplifier that isolates the modulator and associated circuits from the frequency multiplier circuits which follow. It uses cathode bias provided by resistor *R*8 with bypass capacitor *C*6. The output voltage is developed across the tuned plate circuit formed by *L*1 and the distributed capacitance. Screen voltage is applied through *R*9, with *C*7 and *C*8 operating as conventional r-f bypass capacitors. The voltage developed across *L*1 is still at the crystal frequency, but has been increased considerably in amplitude.

f. The voltage across *L*1 is coupled to the grid of the first frequency multiplier, *V*6, through *C*9, and is rectified between the grid and cathode, since its positive swings are sufficient to draw grid current. This develops bias across *R*10, causing the stage to operate in class *C*. The desired harmonic is selected by the tuned plate circuit formed by *C*11 and *L*2, in which *C*10 serves as an r-f bypass capacitor. This higher frequency is coupled to another frequency multiplier, *V*7, identical in operation with *V*6. The components of the stage perform functions equivalent to those of the preceding stage. The tuned plate output circuit of the second multiplier, *V*7, is tuned to the operating frequency desired.

g. Drive from tuned circuit *L*3 and *C*13 is transformer-coupled to the cathode circuit of *V*8, *L*4, and *C*15. This stage supplies excitation for the power amplifier. Both the driver and the power amplifier are grounded-grid stages, as necessitated by the high operating frequency. Bias for the grounded grid is achieved through combination *R*12 and *C*16 in *V*8, and *R*13 and *C*20 in *V*9. Since sufficient voltage is applied to the cathode to make it negative in respect to the grid, current flows in the grid circuit and builds up bias across the resistor. The capacitor acts as a bypass, and effectively grounds the grid for r-f.

h. The plate output circuit of the driver stage formed by *L*5 and *C*17 is coupled to the power-amplifier cathode circuit, *L*6 and *C*19. Output from the entire transmitter is applied to a transmission line by the pi-network impedance-matching circuit formed by *C*21, *L*7, and *C*22. This circuit permits matching a wide range of transmission lines by adjustment of the variable capacitors. Plate voltage is fed on the low-impedance side of the circuit through an r-f choke, RFC, and is prevented from reaching the antenna by blocking capacitor *C*29.

47. Direct F-M Transmitter Circuit

a. The circuit for a direct f-m transmitter generally will be more complicated than most of the indirect types when it incorporates automatic frequency control. If no afc is used, considerable simplification in the number of stages can be obtained at the expense of reduced frequency stability. Because most direct f-m systems are capable of a higher deviation of the fundamental oscillator frequency than are indirect units, less frequency multiplication is needed to attain the desired operating frequency and deviation. A representative direct f-m transmitter with automatic frequency con-

trol is shown in the complete schematic of figure 100. Although the diagram looks extremely complex at first glance, it can be understood easily if it is analyzed stage by stage. The power supply and control circuits have been eliminated for clarity.

b. The circuit of tube *V*1 is a basic Colpitts oscillator. The frequency control network comprising *V*3 and *V*4, along with reactance modulator *V*2, serves to keep it in synchronism with crystal oscillator *V*5. Audio voltage is amplified by *V*13 and applied to the reactance modulator to produce the necessary deviation. The remaining stages in the transmitter are buffer amplifiers, frequency multipliers, and the final output stage.

c. The tank circuit of oscillator *V*1 is formed by *L*1 and the split-feedback capacitors, *C*1 and *C*2. *R*1 and *C*3 are a conventional grid-leak bias combination. An r-f choke is used in the cathode so that d-c plate current can flow from plate to cathode, and r-f currents are diverted to the feedback circuit. A conventional plate tank circuit, *C*6 and *L*2, along with bypass *C*5, is used. This tank circuit is tuned to a harmonic of the oscillator frequency to obtain increased isolation of the frequency-determining components from the output of the oscillator. In addition, a stage of frequency multiplication is saved. Screen voltage is fed to the oscillator tube through the voltage divider formed by *R*2 and *R*37. When the screen voltage is adjusted properly, the oscillator frequency is practically independent of changes in plate voltage.

d. Output from the oscillator is coupled through *C*7 to the grid of buffer amplifier *V*6, which operates class A as determined by the bias built up across *R*4 in the cathode circuit. Capacitor *C*8 acts as a bypass for r-f and *R*3 serves as a d-c grid return. It does not affect the bias because the output of the oscillator is adjusted so that negligible grid current is drawn. The screen voltage is supplied from the main plate supply through *R*5. The screen itself is grounded for r-f by *C*9, and the bottom of the plate tank circuit, *L*3 and *C*58, is bypassed by *C*10. The output of this stage is fed to *V*7 of the first frequency multiplier through *C*11. Bias is developed by grid-cathode rectification of the amplified signal across *R*6. The output of this multiplier appears across the tuned circuit formed by *C*13 and *L*4, which is returned to ground for r-f by bypass *C*12. This output is fed in a similar manner to *V*8 of the second multiplier which in turn feeds *V*9 of the third multiplier. The output of the third multiplier appears across the tuned circuit formed by *C*18 and *L*6, which is bypassed to ground for r-f by *C*19.

e. Examination of the frequency modulator discloses that *V*2 is the conventional injected reactance type, with *C*35 and resistor *R*18 forming the phase-splitting network. Resistor *R*17 is inserted in series with *C*35 to prevent the development of very-high-frequency oscillation, which can be caused by *C*35 in association with stray wiring inductance. This can act as a tuned circuit and permit the development of ultraudion oscillation at frequencies where the screen loses its effectiveness in reducing capacitance between grid and plate. *C*34 is a blocking capacitor that prevents d-c from appearing in the grid circuit. Isolation is increased by bypass capacitor *C*36. The screen voltage and bypassing are supplied by *R*20 and *C*37. Operating bias for the modulator is set by *R*19, and the cathode is grounded for r-f and audio by *C*38.

f. Audio voltage is fed to the modulator grid through isolating resistor *R*16 and coupling capacitor *C*54 from audio amplifier *V*13. The audio-amplifier stage has a pre-emphasis circuit in the grid formed by *C*57, *R*34, and *R*33. *C*57 is selected so that, in combination with *R*33, a voltage divider is formed which presents an increasing voltage with frequency caused by the decreasing reactance of *C*57 as the frequency is raised. To limit the attenuation of low frequencies *R*34 is included. Since it is in parallel with *C*57, the maximum impedance that can be developed is the resistance of *R*34 alone, when the frequency is so low that *C*57 has an extremely high impedance. *R*35 acts as a gain control and as a constant source impedance for the pre-emphasis network.

g. The frequency-control circuit includes *V*3, *V*4, and *V*5. It compares the output voltage of

one of the multiplier stages with a harmonic of the crystal oscillator in a mixer, *V*4. If any difference exists between the two, a voltage is developed in the phase discriminator, *V*3, which is applied to the grid of the frequency modulator through isolation resistor, *R*21. This compensates for the frequency difference. The mixer stage uses a pentagrid tube which permits application to separate grids of the two signals to be compared. The difference frequency is selected by the plate circuit. This usually is tuned so that the entire discriminator operates at a low frequency. The crystal oscillator, *V*5, is a special type which permits the selection of a harmonic of the actual crystal frequency by means of the tuned plate circuit formed by *L*13 and *C*50. The harmonic of the crystal frequency is selected to give a difference between it and the harmonic of the master oscillator appearing at the grid of the driver tube, *V*10. *R*29 acts as the grid-bias resistor for the oscillator, with the crystal itself acting as the grid-leak capacitor. Output is coupled to the mixer tube through *C*47 and *R*26, and the output from the frequency multiplier *V*9 is coupled through *C*48 and *R*27.

h. The output of the frequency multiplier stages feeds a tetrode driver stage whose plate circuit is tuned to the operating frequency. This tuned circuit, *C*23 and *L*7, is grounded in the center for r-f by capacitor *C*22 so that a 180° out-of-phase voltage is available for the grids of the final push-pull amplifier, *V*11 and *V*12. A small variable capacitor, *C*26, is included on the side of the tank opposite the plate of the driver. It compensates for the capacitance from plate to cathode that would tend to unbalance the driving voltages to the grids of the power amplifier. Cathode bias is provided in the amplifier by *R*13 and *R*14 as a safety measure in case excitation fails, but the main operating bias comes from grid rectification and resultant production of voltage across *R*11 and *R*12. The final tank circuit, formed by *C*31, *C*32, *L*8, and *L*9, is coupled to a transmission line by the tuned link circuit. The screen-voltage supply for the final amplifier is conventional and is similar to that used for the driver.

Section IV. TYPICAL TRANSMITTERS

48. Indirect F-M Transmitter

A general description of a transmitter must include characteristics other than circuitry. These include the power input and output, the frequency and type of operation, the purpose of the equipment, and the distance over which the signal from the transmitter can be received. A description of these characteristics is given for the indirect f-m transmitter shown in figure 101. This equipment is designed for fixed-station service over a range of approximately 20 miles, with an output of 50 watts. A 110-volt, 60-cycle power supply is required for the power input, and the equipment draws 325 watts from the line when transmitting. The complete unit contains 28 tubes including rectifiers, and has an audio response from 300 to 4,000 cycles, with a frequency range from 30 to 40 mc. The transmitter unit has been removed from the cabinet and is shown in figure 102. The crystal oscillator is phase-modulated through an audio correction network. The resultant f-m signal is multiplied 32 times and produces a frequency deviation of ±15 kc. This is accomplished by 2 quadrupler stages and a doubler stage. The output of the doubler then is fed to push-pull power amplifiers, and approximately 50 watts of power is delivered to the antenna. The receiver (fig. 101) shown in the rack below the transmitter is a double conversion superheterodyne.

49. Direct F-M Transmitters

a. A combination double-conversion superheterodyne receiver and direct f-m transmitter is shown in figure 103. This set provides two-way phone communication with a similar portable, ground, or mobile equipments over a range up to a mile. Power is supplied by either dry cells, or a vibrator power supply and vehicular

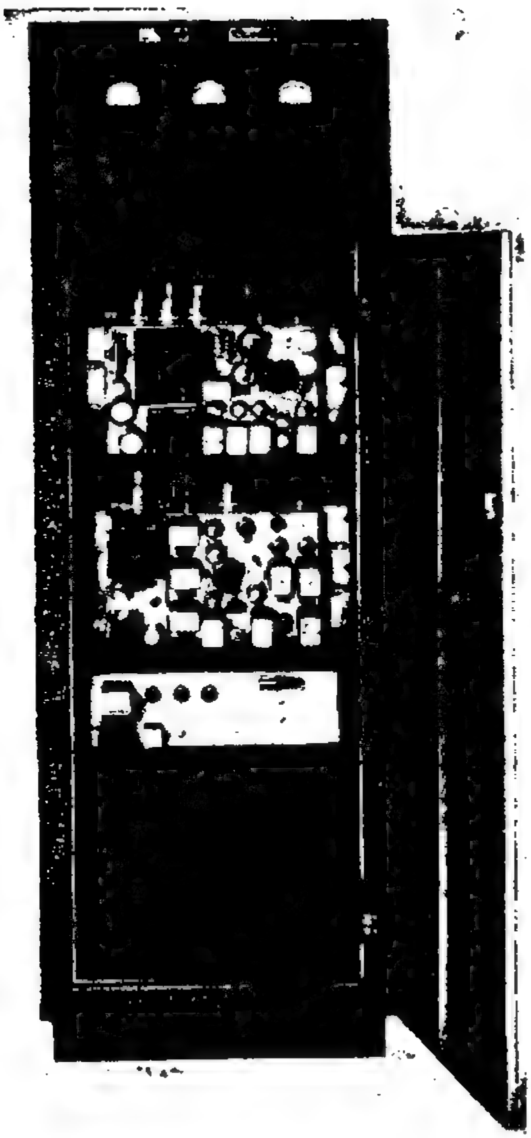


Figure 101. Indirect f-m transmitter used for fixed-station service.

battery. The power supply must provide 90 volts at 80 ma for the plates and screens of the tubes, and 6.3 volts at 520 ma when transmitting. The power output of the transmitter is 500 mw (milliwatts). The frequency deviation is ±20 kc when a 1,000-cycle signal input of .25 volt is applied. The tuning of the transmitter is variable continuously from 47 to 58.4 mc. This band of frequencies provides 115 channels with each channel having a bandwidth of 100 kc. Two preset detented channels are available. The receiver i-f components and the audio components of both the transmitter and receiver are mounted on the i-f chassis, as shown on the right side of figure 104. The view on the left of this figure shows both r-f and i-f components.

b. The transmitting and receiving circuits are associated with each other through a common antenna circuit, a common 32- to 42.3-mc Colpitts oscillator circuit, and a common tuning control. The transmitter converts speech signals from an external microphone, amplifier, telephone line, or other a-f source into f-m signals. The microphone voltage is amplified by a microphone amplifier to the proper value for modulation. It then is applied to a reactance modulator which varies the frequency of the Colpitts oscillator in accordance with the audio signal. The frequency-modulated output of the Colpitts oscillator and a 15-mc signal produced by doubling the low-frequency output of a 7.5-kc crystal oscillator are combined in a mixer stage and the sum frequency of the transmitter mixer is selected by a tuned circuit. The signal then is fed through the driver and power amplifier stages to the antenna. No antenna switching is provided since the receiver is inoperative when the transmitter is energized and the transmitter is inoperative when the receiver is on.

c. Another direct f-m transmitter receiver for portable, ground, or vehicular installation is shown in figure 105. The frequency range of this transmitter receiver is from 20 to 27.9 mc. The communication range is approximately 10 miles for vehicles in motion and 15 miles for stationary vehicles. Power can be supplied by dry cells, vibrator power supplies, or a hand generator. The set contains two subchassis; the i-f chassis mounts the receiver components and the audio components for both transmitter and receiver; the r-f chassis contains the high-frequency parts of both transmitter and receiver. A low-frequency, self-excited, modified Hartley oscillator is modulated by a reactance tube to produce direct f-m. A crystal oscillator controls the frequency of both the transmitter and the

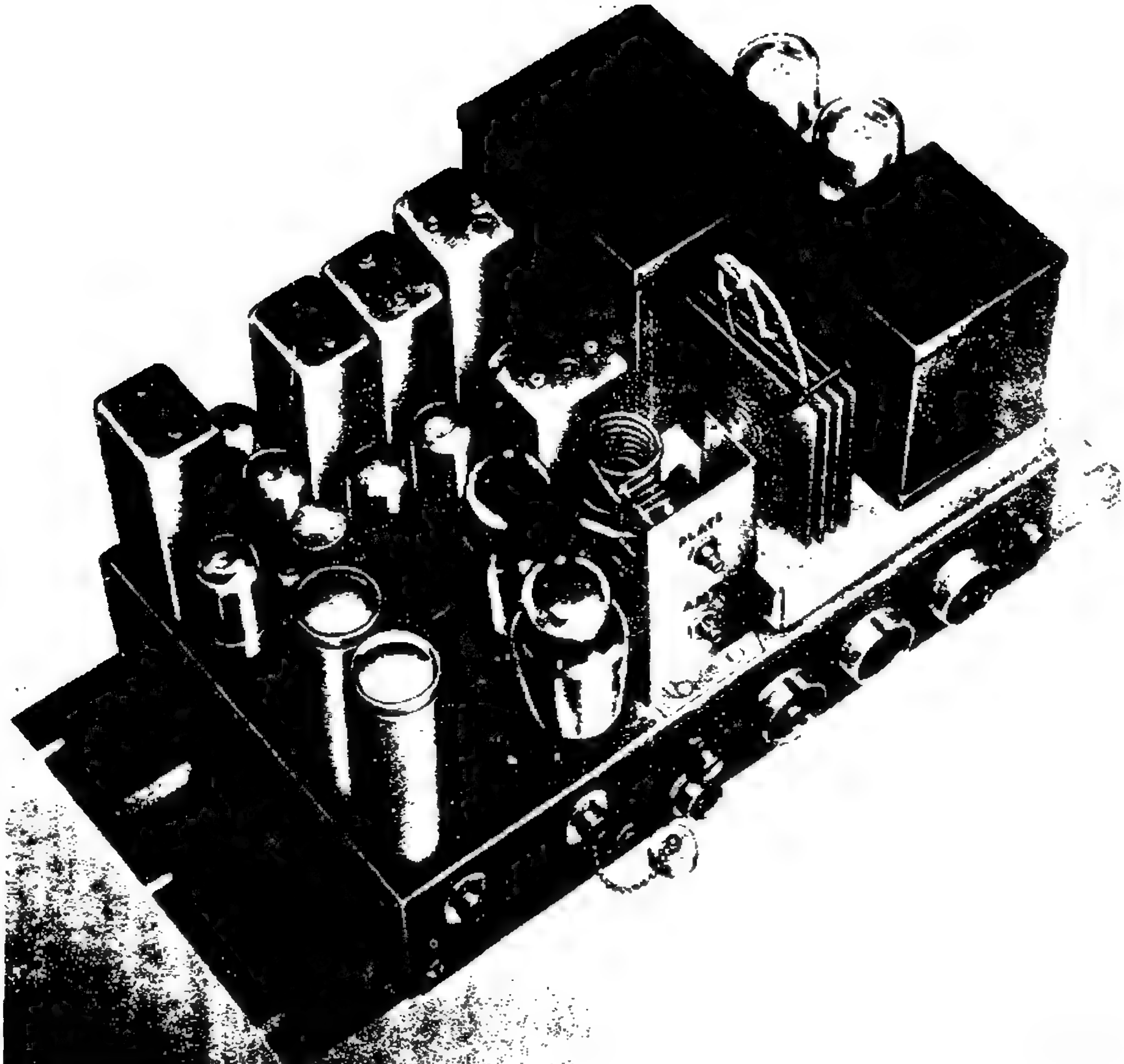
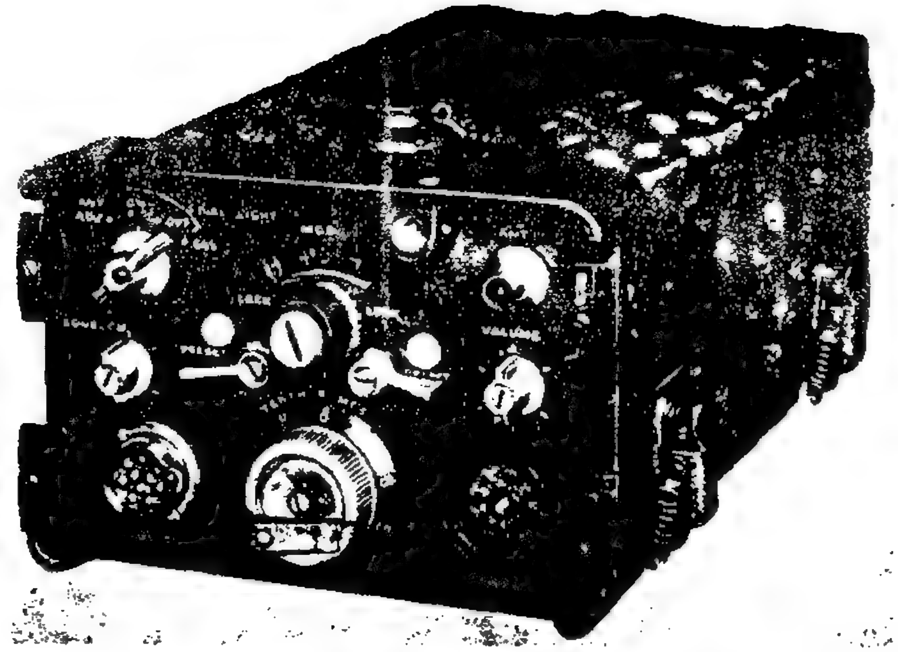


Figure 102. Indirect f-m transmitter removed from rack.

receiver. The output of this oscillator is mixed with the f-m signal to produce the desired deviation of ±20 kc. The output of the mixer then is fed through the driver and power amplifier stages to the antenna.

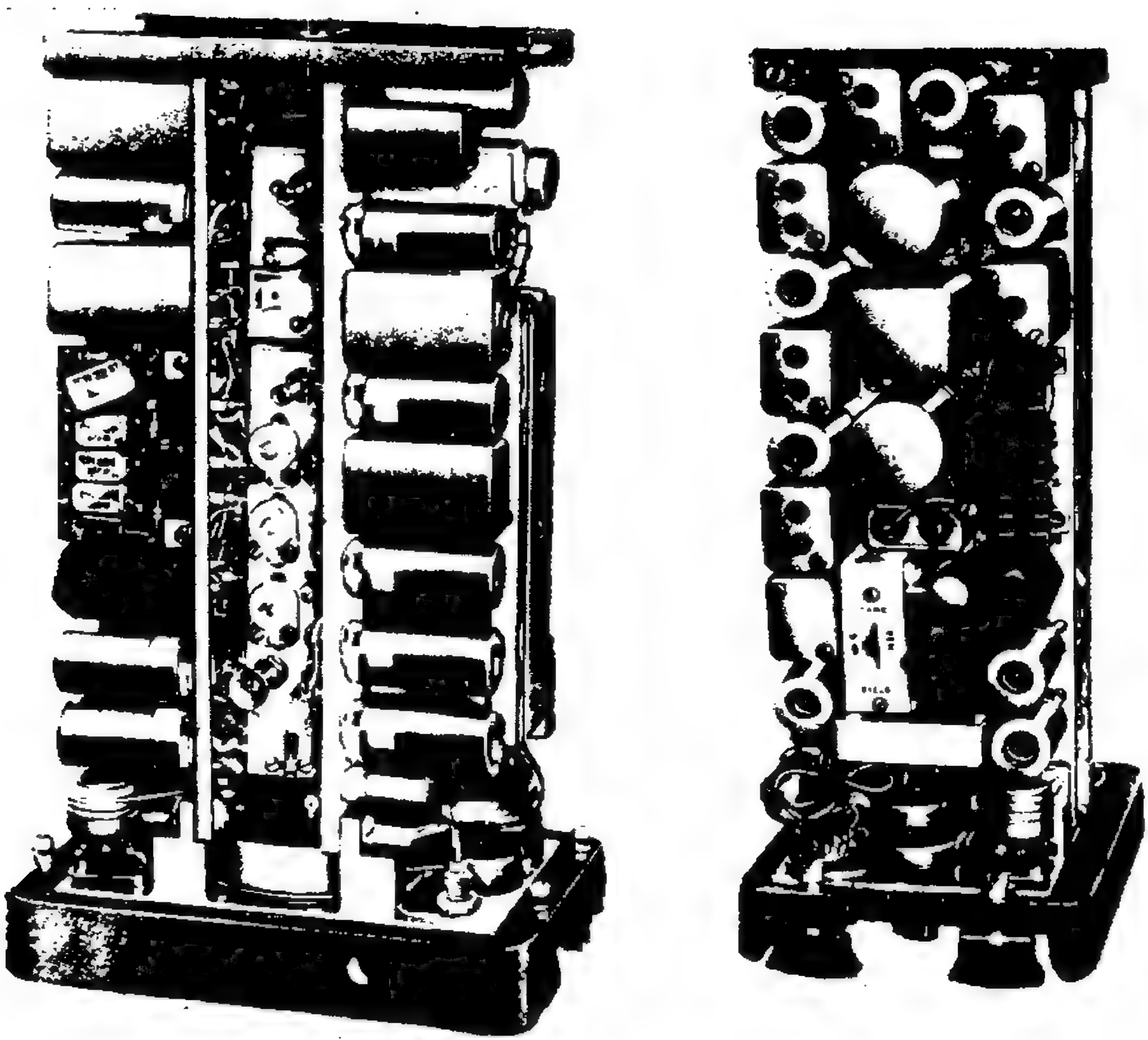
d. A direct f-m receiver transmitter unit designed for pack operation is shown in figure 106. This set has an average distance range of 5 miles. The frequency range is from 27 to 38.9 mc. This band of frequencies provides 120 chan-



TM 668-103

Figure 103. Direct f-m receiver-transmitter.

nels 100 kc wide and the transmitter is preset to any two of these channels. Power can be supplied by dry cells or storage batteries and a dynamotor. The power output of the transmitter is 1½ watts. The entire unit contains 19 tubes, all of which are used when transmitting. The transmitter is a MOPA (master oscillator power-amplifier) with the receiver crystal controlling the frequency of transmission through a reactance tube.



TM 666-104

Figure 104. Direct f-m transmitter-receiver subchassis.

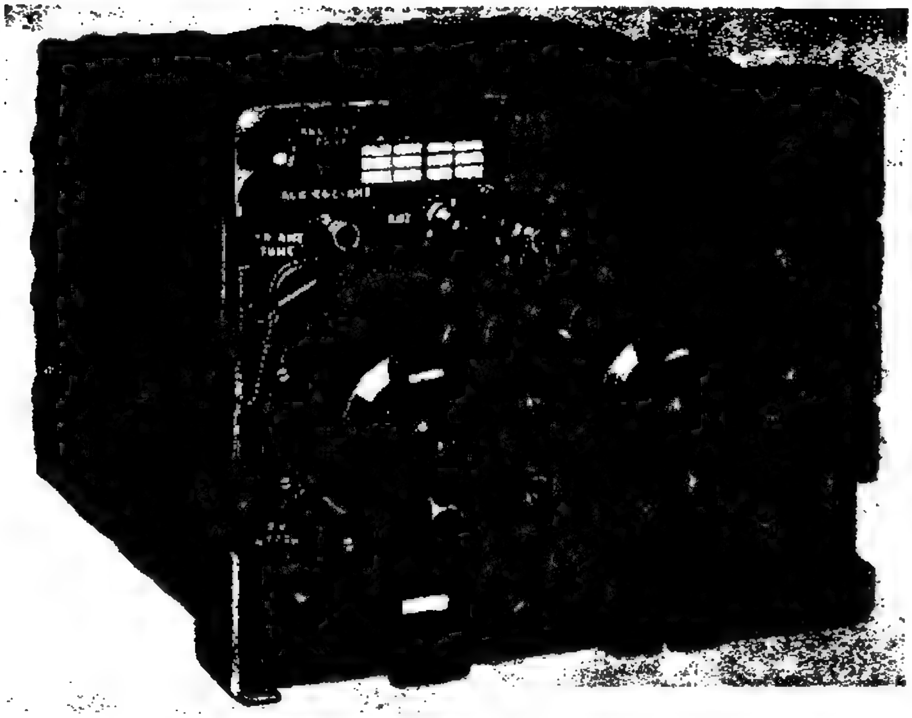


Figure 105. Direct f-m transmitter for vehicular use.

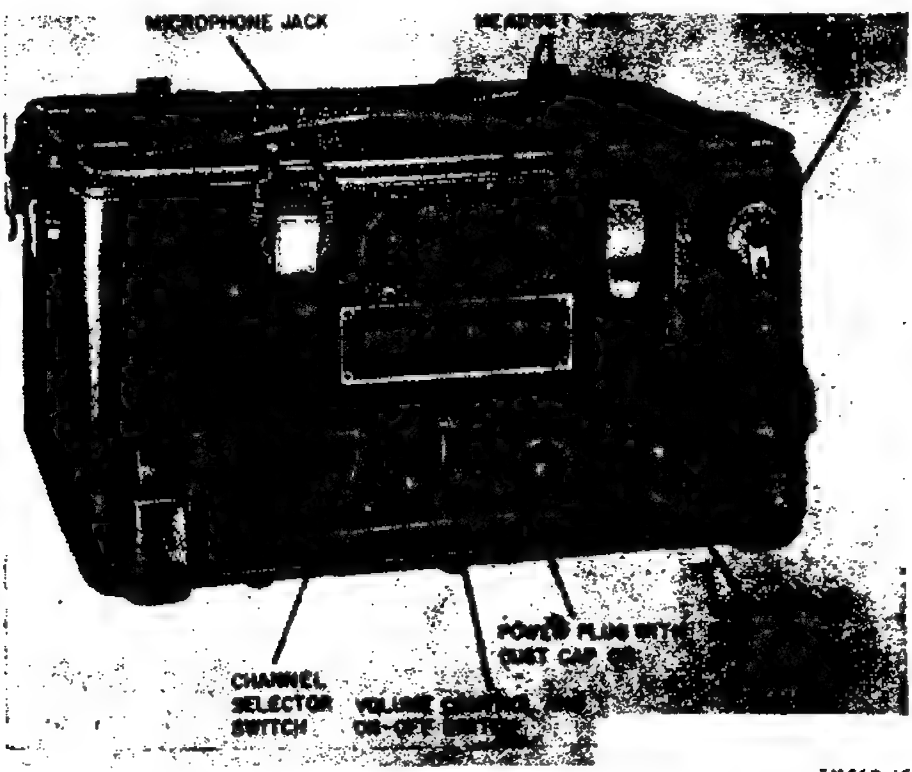


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Figure 106. Direct f-m transmitter for pack operation

Section V. SUMMARY AND REVIEW QUESTIONS

50. Summary

a. The f-m signal taken from the modulator-oscillator stage must be increased in deviation and frequency by frequency-multiplier stages.

b. Frequency multipliers operate through harmonic generation. Multiplications of two, three, four, and five times are obtainable with reasonable efficiency.

c. Frequency multipliers are usually r-f amplifiers with input and output circuits tuned to different frequencies harmonically related to each other.

d. Push-push doublers are very efficient multipliers which use two tubes with grids connected in push-pull and plates in parallel. Operation is similar to a full-wave rectifier producing a ripple voltage of twice the line frequency.

e. At high frequencies, the operation of frequency multipliers is impaired because of degenerative effects which tend to reduce their power output. These degenerative effects can be neutralized with appropriate circuits.

f. Other methods of frequency multiplication are available which do not depend directly on harmonic generation, but these seldom are used.

g. Frequency multipliers can be combined with the master oscillator circuit for compactness and reduction of power consumption.

h. Power amplifiers for f-m have no connection with the modulation process. They are used only to increase the power in the modulated wave. All f-m power amplifiers operate class C, which permits high efficiency and high power output.

i. The operation of a class C amplifier depends on the angle of plate current flow, which in turn depends on the grid bias and the amplitude of the grid-driving voltage.

j. Various class C amplifier circuits have been devised with different means of introducing the grid-driving voltage and coupling the output power to the load.

k. The necessity for neutralization is overcome at high frequencies through the use of tetrode amplifiers.

l. Grounded-grid triode amplifiers overcome some of the disadvantages of tetrodes and conventional triodes at very-high frequencies.

m. The input circuit for a power amplifier must provide adequate regulation of the driving voltage without causing excessive losses. Various input circuits have been devised by use of capacitive coupling or inductive coupling.

n. Power-amplifier output-coupling circuits are of two forms, series-fed and shunt-fed. The output load circuit matches the final amplifier to the transmission line or the antenna. In addition, it suppresses spurious frequencies. Various types of output circuits provide different degrees of suppression and ability to match loads of different impedance.

o. Not all of the various input and output circuits available for class C power amplifiers can be used with each other because of the possibility of spurious oscillations being generated in the circuits.

p. Spurious oscillation generated in a power amplifier at a frequency far from the actual tuning of the tank circuits is called parasitic oscillation. Parasitic oscillations are eliminated by the proper choice of circuit in both grid and plate, as well as by suppressor chokes, resistors, and capacitors.

q. In direct f-m transmitters, the center frequency must be stabilized by some form of automatic frequency control.

r. There are two major types of frequency-control arrangements. One uses a d-c correction voltage for the modulator circuit, the other a mechanical arrangement.

s. Afc systems include frequency changers, comparators, and corrector networks.

t. The comparator is usually a discriminator circuit. The discriminator produces a d-c voltage proportional to the difference between the frequency of the oscillator and some standard center frequency.

u. The double-tuned discriminator uses two resonant circuits separated slightly in frequency and coupled to the same source. The output is applied to rectifiers and the amplitude of the rectified voltage is proportionate to the change in frequency.

v. The phase discriminator compares the phase of signals from the master oscillator and a crystal standard, producing a d-c output proportional to the difference.

w. The pulse discriminator produces pulses whose polarity and number per second reflect the departure of the oscillator from the center frequency. These pulses are stored, and any change in stored charge is applied to the modulator tube as a correction voltage.

x. In a motor control system, two balanced modulators, receiving their signals from the crystal oscillator and master oscillator, produce two outputs 90° out of phase if the signal frequencies do not coincide. The 90° out-of-phase voltages are applied to the windings of a two-phase motor. This motor turns a variable capacitor in the oscillator circuit, which serves to correct the frequency difference.

y. Frequency dividers are of two types—free-running dividers which produce output whether input is present or not, and those which produce an output only with an input signal.

z. The multivibrator is a common free-running divider. It is essentially a two-stage resistance-coupled amplifier with some of the output returned to the input. The essentially square waveform it produces can be synchronized with higher harmonics.

aa. Trigger circuits are modifications of multivibrator circuits which operate only in the presence of input pulses. Used in sequence, they operate as very accurate dividers.

ab. The regenerative modulator consists of a nonlinear device and a frequency selective amplifier. The amplifier is tuned to a subharmonic of the input frequency of the nonlinear mixer. The output of the amplifier is returned to the mixer, reinforcing the subharmonic.

51. Questions

a. What increases the deviation and carrier frequency in an f-m transmitter?

b. What feature of the operation of a class C amplifier permits the successful generation of output power at a harmonic frequency of the input?

c. Cite two ways of obtaining a frequency multiplication of 32.

d. How many times is the deviation of a basic f-m signal increased after passing through four doublers?

e. If the input circuit of a quintupler is tuned to 10 mc, to what frequency is the output circuit tuned?

f. Which type of multiplier chain produces greater output, one containing a doubler and a quadrupler or one containing three doublers?

g. How are the grids and plates of a push-push doubler connected?

h. What type of loading is produced by the capacitance between the grid and the plate circuit of a high-frequency multiplier?

i. What is the effect of a high-cathode inductance on the output capacitance of a high-frequency quadrupler?

j. What is the advantage of using push-pull multipliers?

k. What coupling exists between the input and the output circuit of a frequency multiplier which is combined with an oscillator?

l. Why is it desirable to tune the output circuit of a combined oscillator multiplier to a very much higher frequency than that of the grid circuit?

m. What differences exist in the requirements for a class C power amplifier used for f-m as compared to one used for a-m?

n. What is the principal disadvantage of the grounded-grid amplifier?

o. Why are push-pull amplifiers useful at high frequencies?

p. What is significant about the input impedance of a grounded-grid push-pull amplifier?

q. What are the major disadvantages of capacitive coupling as compared to link coupling?

r. What is necessary in respect to the power transfer between driver and final amplifier?

s. What are the functions of the power-amplifier output coupling network? Which of these features is determined by the operating *Q* of the circuit?

t. What are the relative advantages and disadvantages of shunt feed?

u. What is the function of a split-stator capacitor in a single-ended neutralized triode power amplifier?

v. How does the type of feed affect the d-c voltage appearing across the tank capacitor?

w. What is the major disadvantage of the pi-network as an output coupling circuit?

x. How can short antennas be coupled directly to the final amplifier with good efficiency?

y. What are the major operating indications that a final amplifier is oscillating parasitically?

z. How are parasitic oscillations suppressed?

aa. Why are certain combinations of input and output circuits avoided?

ab. What is the difference between a d-c control system and a mechanical control system?

ac. What are the relative disadvantages of a d-c control system?

ad. Compare the response speeds of a motor-control and a d-c control system. Which is more suitable for long-time changes?

ae. What is a receiver-transmitter frequency interlock system?

af. What device is most frequently used to compare the frequencies of two waves?

ag. What is the major disadvantage of the double-tuned discriminator?

ah. Why are discriminators used at a frequency that is low compared to the carrier frequency?

ai. What is the advantage of the phase discriminator as compared to the double-tuned discriminator?

aj. How does the modified phase discriminator differ from the standard phase discriminator?

ak. What is the advantage of a pulse control system as compared to a simple discriminator d-c control system?

al. What is the function of the modified balanced modulator in the pulse control system?

am. What is the function of the limiter tube in the pulse system?

an. Why is the trigger circuit used in conjunction with the pulse discriminator?

ao. How is the output of the pulse discriminator applied to the control of the master ocillator?

ap. Why is the motor control system restricted to large fixed installations?

aq. What are the three major parts of the motor control system?

ar. What is the purpose of using frequency dividers?

as. What is the difference between a free-running and a triggered divider?

at. What is the major disadvantage of the synchronized oscillator as a frequency divider?

au. How is fractional division obtained in a regenerative modulator?

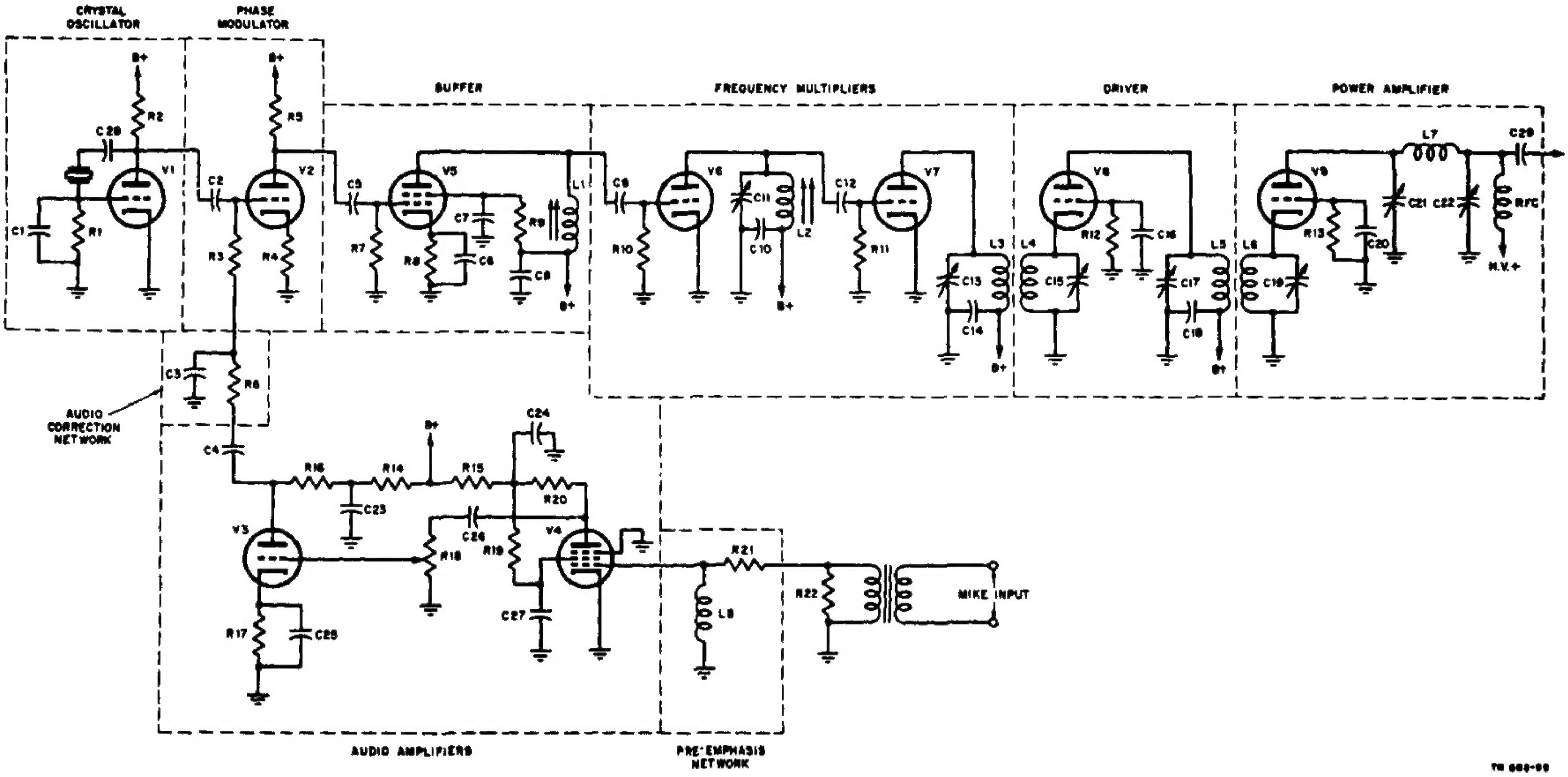

Figure 99. Indirect f-m transmitter.

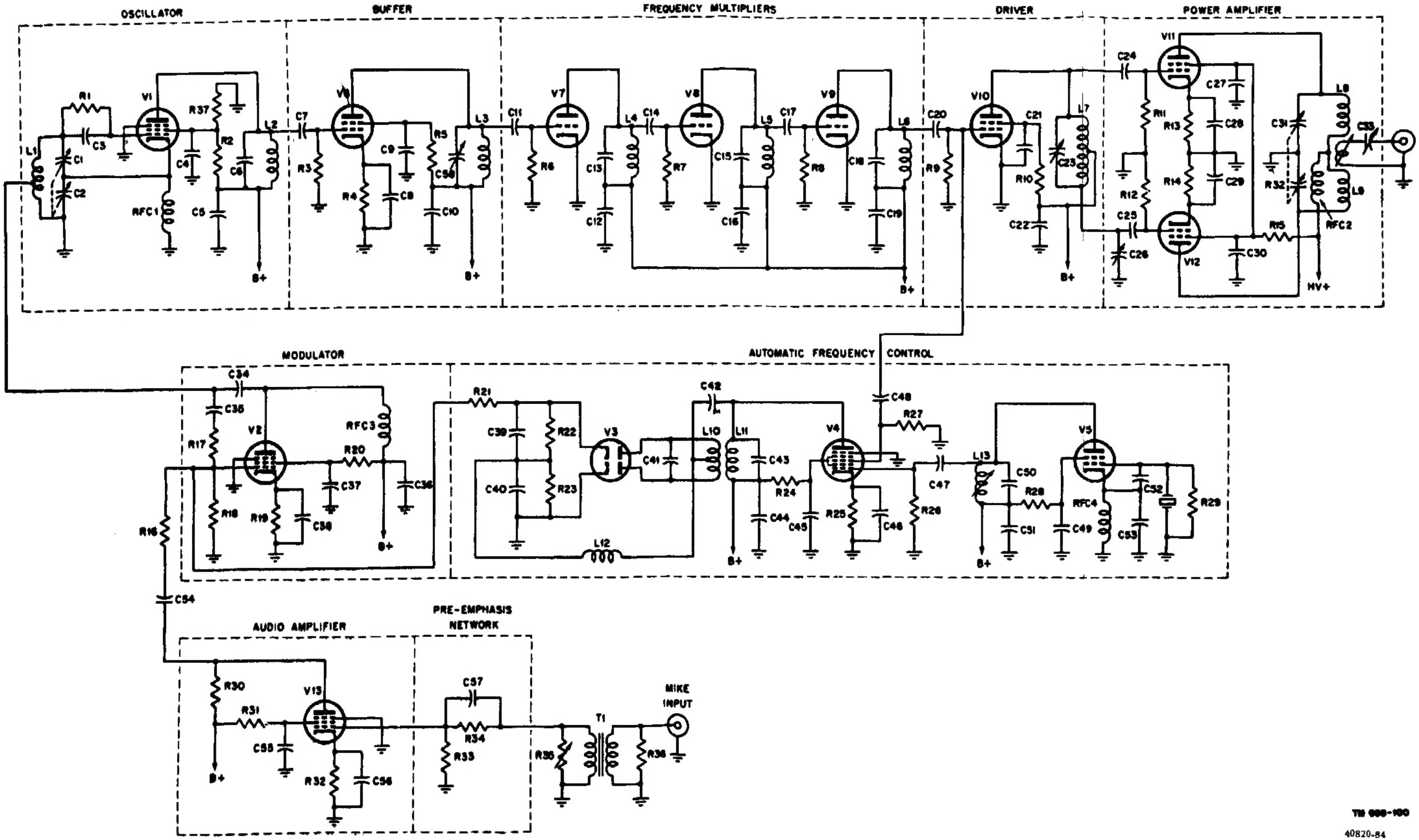

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Figure 100. Direct f-m transmitter.